THE ELECTRONICS MAGAZINE WITH THE PRACTICAL APPROACH

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# Elektor Electronics

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ASIC microcontrollers
Stereo viewer
New generation of analogue switches
High-grade power unit
Simple transmission line experiments
Microprocessor tone generator
Communication receivers front-end filtering

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Elektor Electronics

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September 1989
Volume 15
Number 170

**Theme of the month in October will be Satellite and Cable TV.**

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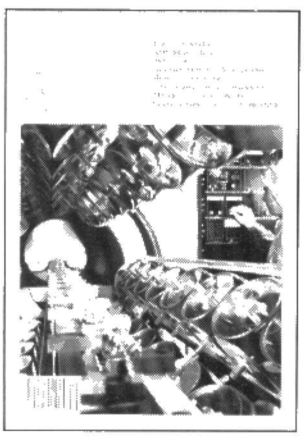
- Inductance meter
- Open systems
- A-D conversion techniques
- CD error detector
- Logic analyser for Atari ST
- Dark-room clock
- Log-antilog amplifier



***Front cover***
Clear compact discs prior to being metallized are seen during production at Numbus Records Ltd, Britain's largest manufacturer, whose development of a new laser-mastering system won the Queen's Award for Technological Achievement in 1987.

Producing a CD master with the Nimbus laser-mastering system involves transferring up to 6,000 million bits of information (recorded sound) on to a prepared glass master. This is then transferred to metal stampers by an electro-forming process. The discs are pressed from clear polycarbonate by fully automated injection moulding presses.

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### General interest



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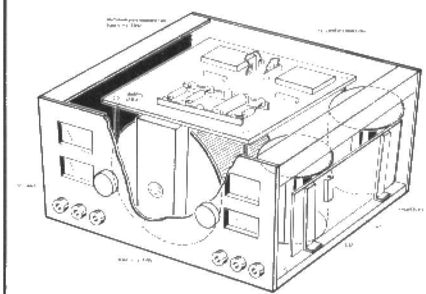

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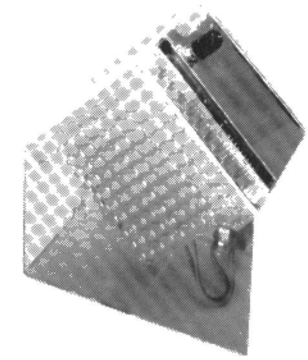

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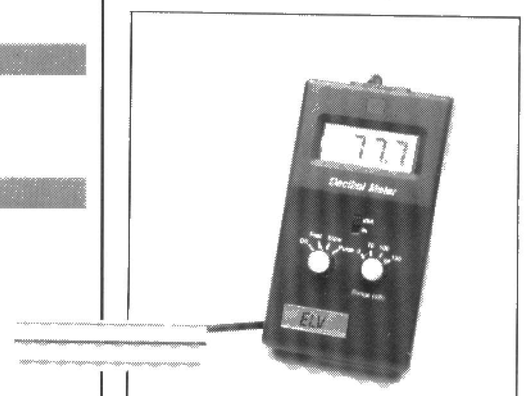


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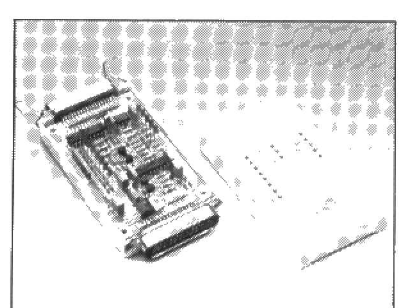

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ELEKTOR ELECTRONICS (Publishing)
Editor/publisher: Len Seymour
Technical Editor: J Buiting
Editorial offices:
Down House
Broomhill Road
LONDON SW18 4JQ
England
Telephone: 01-877 1688 *(National)*
or +44 1877 1688 *(International)*
Telex: 917003 (LPC G)
Fax: 01-874 9153 *(National)*
+44 1874 9153 *(International)*
Advertising: PRB Ltd
3 Wolseley Terrace
CHELTENHAM GL50 1TH
Telephone: (0242) 510760
Fax: (0242) 226626
European offices:
Postbus 75
6190 AB BEEK (L)
The Netherlands
Telephone: +31 4490 89444
Telex: 56617 (elekt nl)
Fax: +31 4490 70161
Managing Director: M.M.J. Landman
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Editor: E J A Krempelsauer
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Elektor sarl
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59270 Bailleul
Editors: D R S Meyer;
G C P Raedersdorf
Greece
Elektor EPE
Karaiskaki 14
16673 Voula – Athens
Editor: E Xanthoulis
India
Elektor Electronics PVT Ltd.
Chhotani Building
52 C, Proctor Road, Grant Road (E)
Bombay 400 007
Editor: Surendra Iyer
Netherlands
Elektuur B.V.
Peter Treckpoelstraat 2-4
6191 VK Beek
Editor: P E L Kersemakers
Pakistan
Electro-shop
35 Naseem Plaza
Lasbella Chawk
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Editor: Zain Ahmed
Portugal
Ferreira & Bento Lda.
R.D. Estefânia, 32-1°
1000 Lisboa
Editor: Jeremias Sequeira
Spain
Ingelek S.A.
Plaza República Ecuador
2-28016 Madrid
Sweden
Electronic Press AB
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# CAN BRITAIN MAINTAIN ITS LEAD IN MOBILE RADIO?

One of the great success stories in the United Kingdom electronics industry over the past few years has been, and still is, mobile radio. Britain's world lead in this field has helped to push the number of two-way mobile radios in the UK to well over a million. Cellular radios, although they became available only in 1985, already account for over half that total. And demand keeps growing.

But while demand continues to grow, there are increasing shortages of skilled engineers and technicians to produce, install and service the equipment. According to the Federation of Communication Services (FCS), the mobile radio market is growing at well over 10 per cent per year, while a survey of its members shows that 90 per cent of them need more technical staff. One industry source estimates that 6,000 more specialist staff will be needed by 1995.

There are several initiatives that mobile radio firms should make the most of to demonstrate that mobile radio offers excellent career prospects. These include the Enterprise and Education Initiative, which aims to strengthen the partnership between business and education by offering young persons the opportunity of gaining work experience in both manufacturing and service industries, and teachers the chance to experience business first hand. Another effective way of drawing school-leavers' attention to the radio industry is through the Young Radio Amateur of the Year Award. This is aimed at anyone under 18 who is keen on DIY radioconstruction or operation, uses radio for a community service, or is involved in amateur radio in some other way, for instance, a school science project.

One of the main problems in mobile radio is the lack of nationally recognized qualifications for technicians. Trainees are often attracted to other sectors where structured training exists. The mobile radio industry itself faces difficulties when recruiting technicians of indeterminate abilities. Consequently, mobile radio users suffer because of the varying quality of service they receive.

These problems led the Department of Trade and Industry, the Mobile Radio Users Association (MRUA), the Federation of Communications Services and the Electronics Engineering Association (EEA) to start the Radiocommunications Quality Assurance Scheme. For a company to maintain certification with the scheme, technicians must be properly trained and qualified. Recognizing this, the DTI and the MRUA earlier this year launched a joint initiative. This resulted in the Mobile Radio Training Committee (MRTC), whose aim is the identification of the mobile radio community's education and training needs.

The dialogue between academics and industry is important. Academics have expressed the view that business people should participate in planning courses and helping to provide on-the-job experience. Educators and trainers should be up-dated by working with businesses, having contact with senior engineers and experiencing the use of modern equipment.

The activities of the DTI, the MRUA, the FCS and the MRTC are drawing attention to the importance and growth of mobile radio. The United Kingdom currently has a leading role, but this position is threatened by the shortage of skilled personnel.

The Government is doing much to highlight the career opportunities and alleviate the problems, but the onus must be on business to form a partnership with education. Packages are required that will attract the people needed, in the numbers required, and provide them with the necessary skills.

# A HIGH-GRADE POWER UNIT

C. Bolton BSc

**These supplies were developed to power experimental electronic equipment including small RF oscillators and amplifiers. There are two versions: a single-channel unit and a dual-channel unit in which the channels may be used independently or in series to give well-balanced positive and negative rails.**

Various circuits may be used to produce a variable, regulated output voltage:

### Chopper circuits

In these, the current is chopped into pulses which are fed to an energy storage device to give an output voltage. This type of circuit was discounted for the present design since the switching involved produces RF energy which readily interferes with other equipment.

### Shunt regulators

These circuits produce a larger current than is required, and shunt the unwanted part away. The shunt regulator is particularly wasteful when the required current is much smaller than the available current, as is frequently the case in experimental work.

### Series regulators

These are in essence series resistors that can be varied to maintain a constant output voltage. Their inefficiency is highest at low output voltage settings and high load currents. Since the ability to power such a load was considered to be the least frequent requirement, this type of circuit was chosen for the design.

**Table 1.**

**High-grade power supply**

Measured performance of each channel:

| | |
|---|---|
| Output voltage: | 0–25 V fully variable |
| Output current: | 1.5 A max. |
| Current limit: | 50 mA–1.5 A |
| Output resistance: | 2 mΩ |
| Output change for 10% mains change: | <1 mV |
| Total ripple plus noise | <1 mV |

Dual-channel unit only:

Output balance: within 10 mV, retained under current limit conditions

The measured performance of the units is summarized in Table 1.

## Single-channel unit

The circuit diagram of the single-channel power unit is given in Fig. 1. The output current is produced by $Tr_2$, $Br_1$ and $C_2$. The output voltage is controlled by a series regulator in which $T_5$, $T_4$ and $T_7$ are the active elements. In effect, these transistors form a multi-stage emitter follower that is driven by opamp $IC_1$. The current gain and the use of Darlington-type power transistors for $T_4$ and $T_7$ ensure a small current demand on $IC_1$. Transistors $T_4$ and $T_7$ are connected in parallel with small emitter resistors to distribute heat dissipation.

The output voltage of $IC_1$ is determined initially by a reference voltage applied to its non-inverting input. The

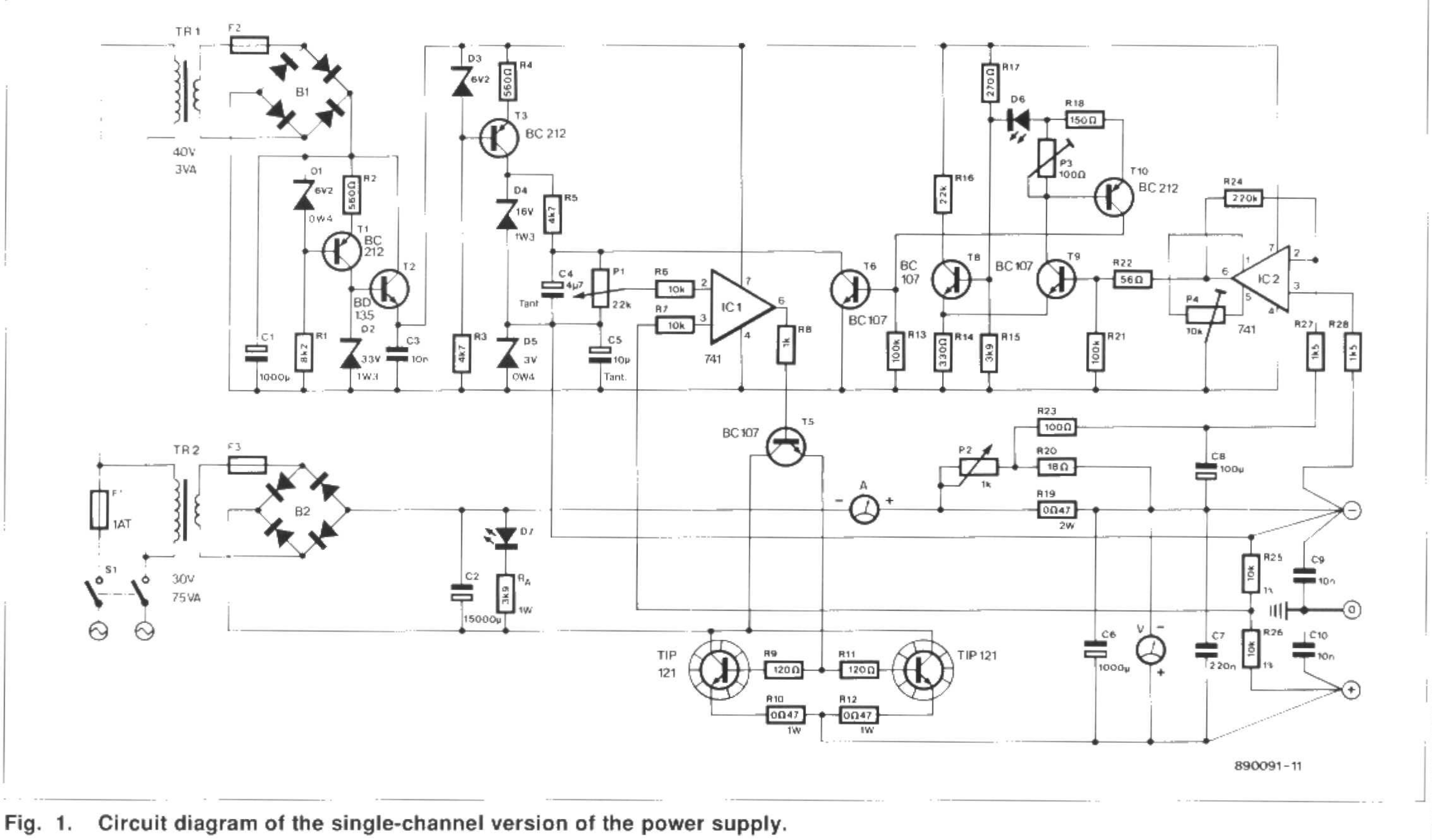


**Fig. 1. Circuit diagram of the single-channel version of the power supply.**

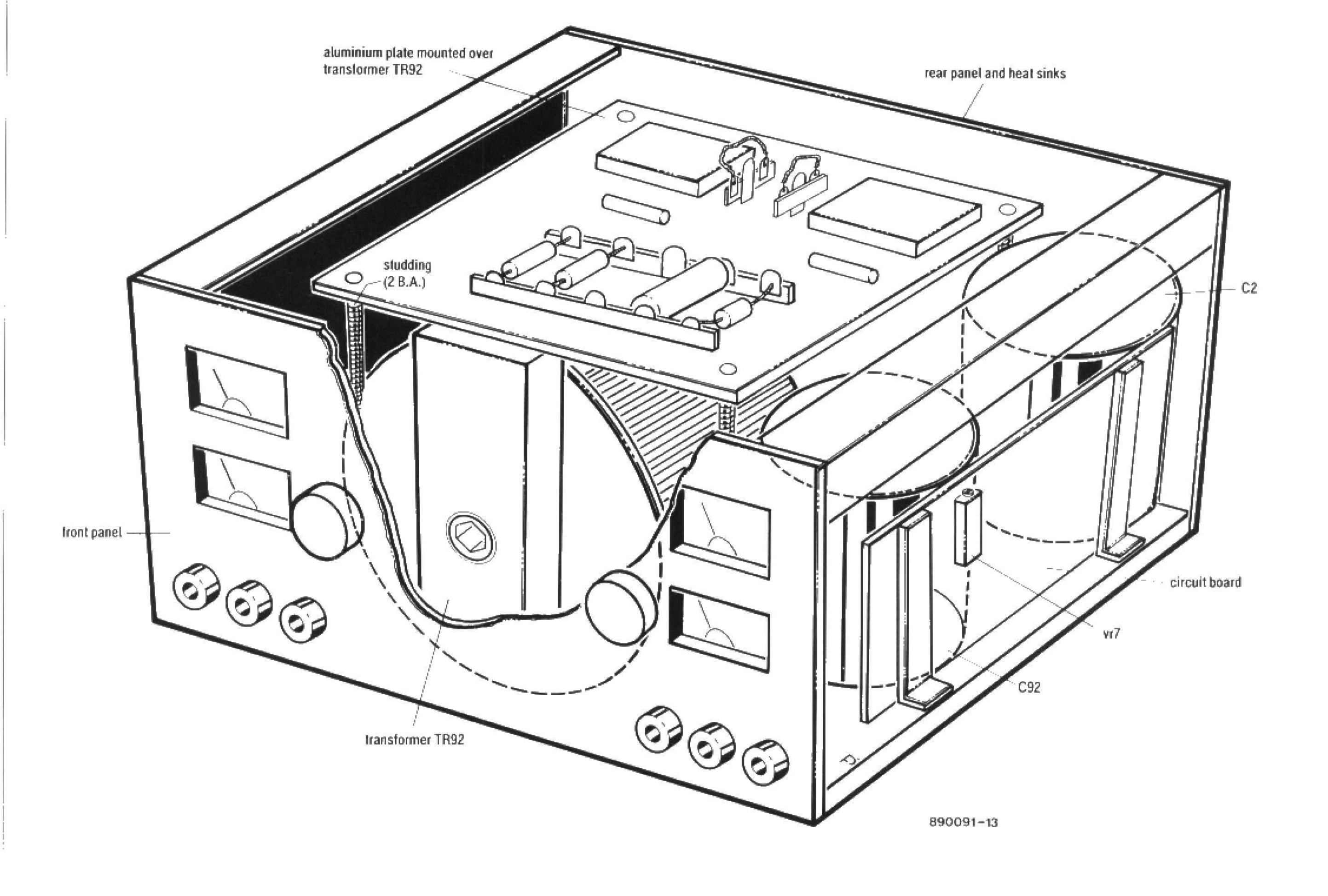


**Fig. 2. Suggested construction of the high-grade power unit.**

inverting input is a fixed fraction of the output voltage supplied by the unit. The high gain and differential operation of $IC_1$ enable the device to vary its output voltage such that the voltage difference between its inputs is almost zero.

The reference voltage for $IC_1$ is derived from constant-current source $T_3$ and zener diode $D_4$. Components $R_5$ and $C_4$ form a simple noise filter. Zener diode $D_5$ produces a small off-set voltage to enable the output voltage to go down to zero.

Current limiting on the basis of voltage feedback is achieved by $R_{19}$, $IC_2$, $T_9$, $T_{10}$ and $T_6$. The current flow through $R_{19}$ produces a voltage across the resistor. Part of this voltage is selected by potential divider $P_2$ and $R_{20}$, amplified by $IC_2$ and applied to a trigger circuit around $T_8$ and $T_9$. Normally $T_9$ is off, but it is turned on when the output of $IC_2$ rises sufficiently because of a higher load current. This causes LED $D_6$ to light, indicating current limiting activity, and $T_{10}$ to be switched on. Transistors $T_9$, $T_{10}$ and $T_6$ now act as an amplifier to draw current through $R_5$, which in turn reduces the reference voltage to $IC_1$ and, consequently, the output voltage.

Power to operate the reference source and associated circuitry is obtained independently of the output supply from $Tr_1$, $Br_1$ and $C_1$, together with stabilizing circuit $T_1$ and $T_2$. Loading on this supply is constant until current limiting occurs. The regulator for this supply thus acts only against fluctuations on the mains, which rarely reach 10%. This enables a steady reference to be obtained fairly simply.

## Practical points

Component layout is not critical, but attention must be paid to a number of points. The can of $C_2$ must be well sleeved to keep it insulated from the chassis. The wires carrying the output voltage must be routed such that they do not form loops enclosing other components (this is most likely to happen on the front panel). On a similar note, the wires carrying the output current must be thick enough to prevent undue heating, and the wire connections at the output terminals must be made exactly as shown in the circuit diagram.

As to cooling, $T_4$ and $T_7$ must be mounted on a heat-sink with a thermal specification not exceeding 0.5 K/W. Remember to insulate these transistors electrically from the heat-sink. Transistor $T_2$ requires only a small heat-sink.

Fuses $F_2$ and $F_3$ are intended to protect the rectifier bridges and the transformers against failure of the smoothing capacitor. They consist of short lengths of 40 SWG copper wire: $F_2$ between vero-pins on the circuit board, and $F_3$ between tags on a short length of tag strip, which can be mounted anywhere convenient to the transformer leads.

Any type of non-steel cabinet may be used to house the power supply. Steel may be used provided the main transformer has sufficiently small magnetic leakage to avoid magnetizing the steel near the leads to the inputs of $IC_2$.

Constructional details of a cabinet that may be made from aluminium are given in Fig. 2. No dimensions are given since these will depend on the components used for $Tr_2$, $C_2$ and the heat-sink. The L-section is extruded aluminium, which is available from many DIY suppliers.

## Setting up

The setting up procedure is concerned entirely with the current limit facility.

1. With the unit switched off, set the output voltage control, $P_1$, for zero volts, the current limit control, $P_2$, for maximum current (maximum resistance), and $P_3$ to zero resistance.
2. Connect a resistor of about 10 Ω, capable of carrying 1.5 A, across the output terminals (a length of electric fire spiral has been found useful).
3. Switch on the unit and raise the output until a current of 1.5 A flows.
4. Adjust $P_4$ so that the current limit warning light, $D_6$, is just on.
5. Increase the resistance of $P_3$ until the

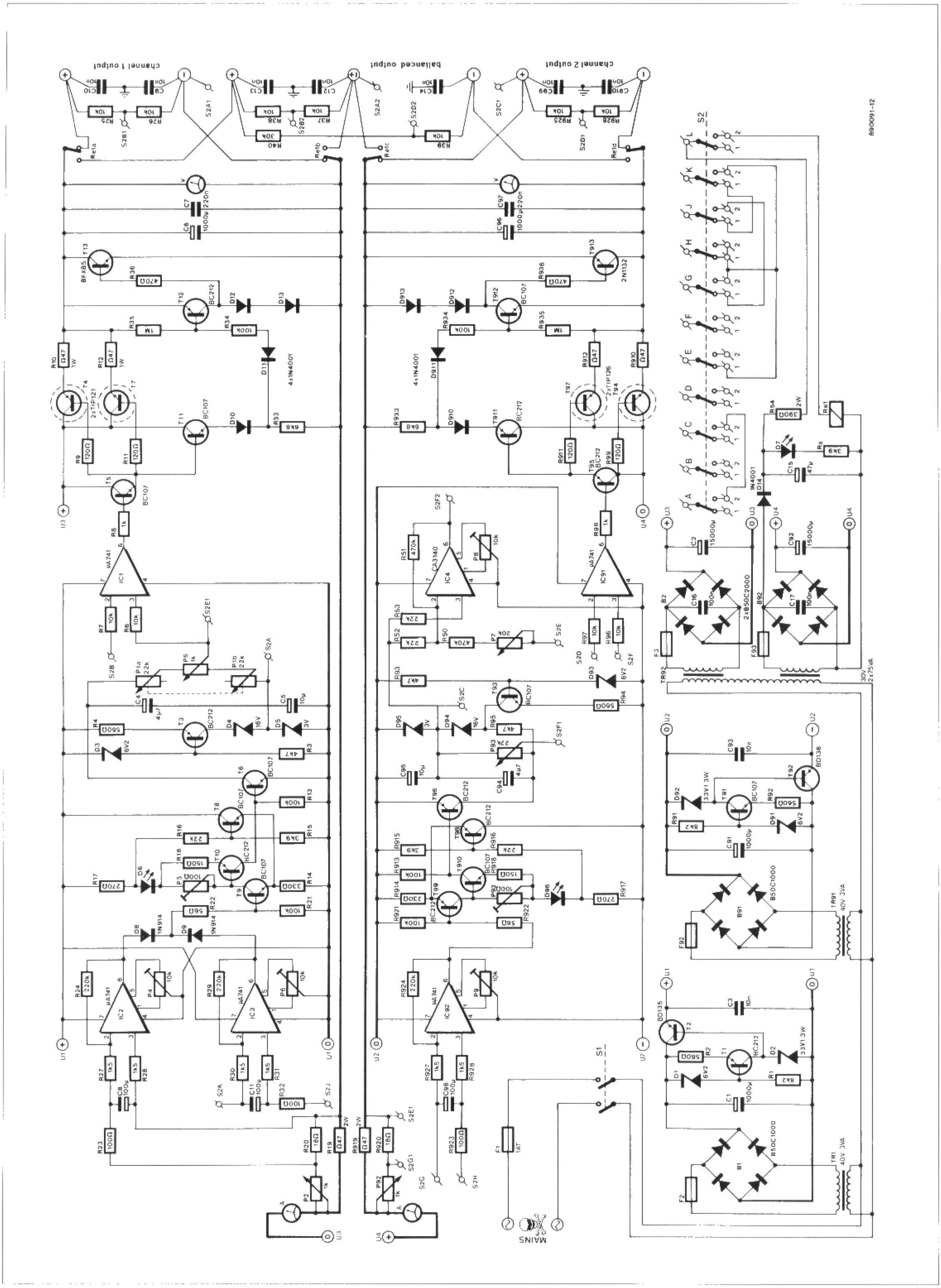

Fig. 3. Circuit diagram of the dual-channel PSU.

current drops by between 50 and 100 mA as indicated on an ammeter.

6. Reduce the output voltage to zero and check that the warning light goes out.

7. Raise the output voltage and check that the warning lamp comes on at 1.5 A.

8. Try to raise the current by raising the output voltage or reducing the load resistor, and check that there is little rise in output current.

9. Choose other settings of the current limit control, and check that limiting occurs at lower currents. The lower limit should be between 30 and 50 mA.

10. If at any time the current limit indicator lights, but at less than full brightness, the limit circuit oscillates because $P_3$ has been advanced too far and should be re-adjusted. This is best done by reducing its value to zero and repeating operations 3, 4 and 5.

The current limit control can be calibrated by setting it to maximum, adjusting the output current to a value required as a calibration point, and then adjusting the limit control until limiting just occurs as indicated by the lamp coming on.

## Notes on the use

The output of the single-channel unit is floating so that either side, or none, may be grounded. The high degree of regulation is available only direct at the output terminals of the supply: bear in mind that six inches of ordinary connecting wire have a higher resistance than the output resistance of the unit.

Under near short-circuit conditions, the current limit may produce a low-level oscillation on the output voltage. This is dependent on the reactance of the load, and is unlikely to be of any consequence since the supply is not normally used as a constant-current source.

## Dual-channel unit

In the dual-channel unit, channel 1 is essentially the same as the single-channel unit. The modifications are the addition of a fine voltage control, $P_5$, a second current limit amplifier, $IC_3$, which is operational only in the balanced mode, and a discharge circuit, $T_{11}$, $T_{12}$ and $T_{13}$, which discharges $C_6$ when the voltage setting is reduced, enabling the output to follow the setting closely.

Channel 2 is similar to channel 1 except that it is complementary. For ease of following, the circuit components with an identical function in channel 1 and the single-channel version are given the same reference numbers. Likewise, components in channel 2 serving the same function as those in channel 1 are given the same numbers with prefix '9'. Thus, $IC_1$ of channel 1 becomes $IC_{91}$ of channel 2. The complete circuit diagram of the dual-channel power supply is given in Fig. 3.

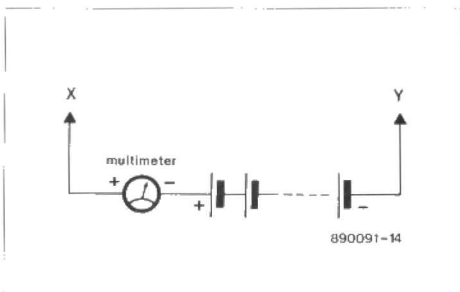


**Fig. 4. Auxiliary circuit for adjusting the PSU.**

## The discharge circuit

As the voltage setting is reduced, the output of $IC_1$ falls, and will fall below the output terminal voltage unless $C_6$ is discharged. The output voltage of $IC_1$ is developed across $R_{33}$ via emitter followers $T_5$ and $T_{11}$. Diode $D_{10}$ produces a small voltage to compensate for additional base-emitter drop in Darlington transistors $T_4$ and $T_7$. If the voltage across $R_{33}$ is lower than that across $C_6$, $T_{12}$ is turned on. This in turn switches $T_{13}$ on, which discharges $C_6$ until the voltage across it is almost equal to that across $R_{33}$ when $T_{12}$ is turned off. Components $D_{12}$, $D_{13}$ and $R_{36}$ limit the base current in $T_{13}$ to a safe level. Diode $D_{11}$ prevents the base of $T_{12}$ being driven dangerously positive if the voltage setting is raised suddenly.

## Balanced output mode

In the balanced output mode, the operation of channel 1 is unchanged. In channel 2, the reference voltage is obtained from the channel 1 reference via the 'times-1' amplifier, $IC_4$. This reference is compared with ¾ of the voltage between the positive and negative rails produced by potential divider $R_{39}$-$R_{40}$.

If the current in channel 1 exceeds the set limit, $IC_2$ causes the limit circuit to operate. If the current in channel 2 exceeds the limit setting, the output from $IC_3$ causes the limiter in channel 1 to operate. Since both channels use the channel 1 reference, they are limited equally in both cases. Hence, balance is maintained under current limiting conditions. Diodes $D_8$ and $D_9$ prevent competition for limiting between the channels. The current limit settings of the two channels are independent.

Switching between modes of operation is accomplished by $S_2$, which is a wafer switch made up of two 6-pole, 2-way wafers. Relay $Re_1$ is operated by $S_1$ to switch the output current.

Current for the relay coil is obtained from the channel 2 AC supply via $D_{14}$ and $C_{15}$. This supply also feeds $D_7$, the 'power-on' indicator.

## Setting up procedure

With the unit set for independent channel operation, set the current limit circuits as described for the single-channel unit. Use $P_3$ and $P_4$ for channel 1, and $P_{93}$ and $P_{94}$ for channel 2.

To adjust the balance, either a digital voltmeter capable of resolving millivolts at 25 V and below, or the auxiliary test circuit shown in Fig. 4, is required. The PSU must be switched on at least five minutes before the balance is adjusted.

Turn $S_2$ to balanced operation. Set channel 1 to about 10 V and adjust $P_7$ so that channel 2, now the negative rail, also supplies 10 V.

**Setting up with a DVM**

11. Set the output voltage to about 19 V with the aid of the channel 1 control.

12. Connect the digital meter to the + and ± terminals. Note the reading.

13. Connect the digital meter to the ± and – terminals. Adjust $P_7$ to give the reading obtained in step 12.

14. Reconnect the digital meter to the + and ± terminals. Reduce the output voltage to 1.5 V and note the exact reading.

15. Connect the digital meter to the ± and – terminals. Adjust $P_8$ to give the reading obtained in step 14.

**Setting up with the auxiliary test circuit**

11. With the 18 V battery in the test circuit and the multimeter on the 25 V range, connect point X to the + terminal, and point Y to the ± terminal. Adjust the output voltage so that the multimeter reads zero. Change to 100 mV and adjust the output voltage to give a multimeter reading of 50 mV.

12. Set the multimeter to the 5 V range. Connect X to the ± terminal, and Y to the – terminal. Adjust $P_7$ until the multimeter reads zero. Change the multimeter range to 100 mV and adjust $P_7$ to give a reading of 50 mV.

13. Disconnect the test circuit from the unit. Replace the 18 V battery by the 1.5 V cell in the test circuit. Reduce the output to about 2 V and set the multimeter to the 5 V range. Connect X to the + terminal, and Y to the ± terminal. Adjust the unit until the multimeter reads zero. Change the multimeter range to 100 mV and adjust the output voltage to give a reading of 50 mV.

14. Set the multimeter to the 5 V range. Connect X to the ± terminal, and Y to the – terminal. Adjust $P_8$ so that the multimeter reads zero. Change the multimeter range to 100 mV and adjust $P_8$ to give a reading of 50 mV.

**Further settings common to both methods:**

16. Repeat steps 11 to 15.

17. Repeat steps 11, 12 and 13. If any adjustment of $P_7$ is required, steps 14 and 15 must be repeated, followed by steps 11, 12, and 13 and so on until no further adjustment is required.

18. Connect the 10 Ω resistor used for setting the current limit to the ± and – terminals. Set the channel 2 current limit control for maximum current, and adjust $P_6$ so that the channel 1 limit warning lamp just comes on when the current in channel 2 (the negative rail) reaches 1.5 A.

# COMMUNICATION RECEIVER FRONT-END FILTERING

by A. B. Bradshaw

In communication receivers, whether intended for general coverage or for amateur bands only, front-end design has changed considerably over the years. With the use of higher intermediate frequencies (IF) and the availability of high-frequency (HF) crystal filters, we no longer see the multiple banks of tuned circuits and multigang capacitors.

Unfortunately, for most new developments there is a price to pay: the reduction in pre-mixer selectivity means that any amplifier preceding the mixer must offer superlative performance in terms of intermodulation distortion and cross modulation. If it does not, the user may get the impression that the receiver is full of signals. The old saying that "The wider the window's open, the more muck blows in" is very apt here.

It was with these thoughts in mind that the writer has designed some general-purpose front-end filters for the amateur bands. If you need more protection up front when Joe Bloggs just down the road fires up his 400 watts of sideband, these filters should help you to listen on the next adjacent band up or down. You may wish to incorporate them in your next receiver.

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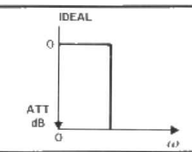
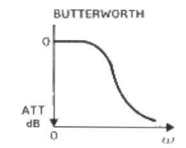
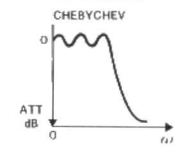
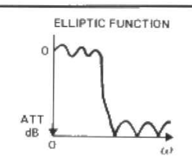
|  | |
|  | Very flat pass band. Attenuation continues to increase in stop band. Simple to set up. Very poor pass–stop band transition. |
|  | Equally spaced ripples in pass band. Attenuation continues to increase in stop band. More difficult to set up than Butterworth. Good pass–stop band transition. |
|  | Unequally spaced ripples in pass band. Attenuation (minimum) is defined in stop band. Ripples in stop band. More difficult to set up than Butterworth. Very good pass–stop band transition. |

**Fig. 1. Frequency response characteristics of the ideal low-pass filter and three approximations.**

## What kind of filter?

Frequency filters fall into four categories: low-pass (LP), high-pass (HP), band-pass (BP), and band-stop.

The design of a band-pass filter for relatively small bandwidths is not too difficult, but the difficulty increases exponentially with increasing bandwidth!

Band-pass and band-stop filters may be constructed from a mix of low-pass and high-pass sections. In BP filters, these sections are in series: in band-stop filters they are in parallel. The band-stop filter so constructed is not often seen in print, but is, nevertheless, a thoroughly practical design. It is, of course, a pity that the LP and HP sections can be used only for the construction of a band-pass or a band-stop filter, but not for both simultaneously!

In modern filter design, a number of approximations to the ideal brickwall response have become popular. The low-pass responses of these are shown in Fig. 1. Their high-pass response is obtained by network transformation.

The operating impedance, band edge, attenuation in the stop band, pass-band ripple and component values are all derived from a low-pass section normalized for a frequency of 1 radian and an impedance of 1 Ω.

The shape of the response, which determines the complexity (length) of the finished filter, is decided with the aid of design tables. There is usually a trade-off between the ratio of the band-edge frequency to the design attenuation frequency and the stop-band attenuation. This means that the 'squarer' the response of a given filter is, the lower will be the stop-band attenuation.

In the construction of a BP filter, the band edge of the LP section becomes the upper profile and that of the HP section, the lower profile. In effect, the two responses cross over each other.

In the designs illustrated in this article, the elliptic function approximation is used. With this, the minimum stop-band attenuation remains at its design figure, in contrast to Butterworth or Chebishev functions where it increases the further the frequency is away from the band edge. This is, however, a small price to pay for the excellent transition band selectivity of this type of filter.

The filters discussed here are designed for a stop-band attenuation of 40 dB or 80 dB to meet both light and stringent requirements. Also, they are designed for an input and output impedance of 50 Ω. Although intended primarily for receiver applications, they may, of course, be used in transmitters, in which case the component ratings MUST BE UPGRADED!

## Components

Ideally, the filters should be constructed on a printed-circuit board, but this is not essential.

Capacitors should be low-loss types. They should be connected in parallel to get their tolerance within 1%, although silver-mica capacitors with 1% tolerance are readily available.

The Q of the inductors should be as high as can be obtained. However, as the filter impedance is 50 Ω, the values of inductance are low, so the coils can be wound manually quite easily. Take care to prevent inductive coupling between sections.

When setting up the filter, trim the coils to their correct value by checking the stop-band nulls on an oscilloscope (or analyser if you are that lucky!). Check all frequencies with a suitable counter.

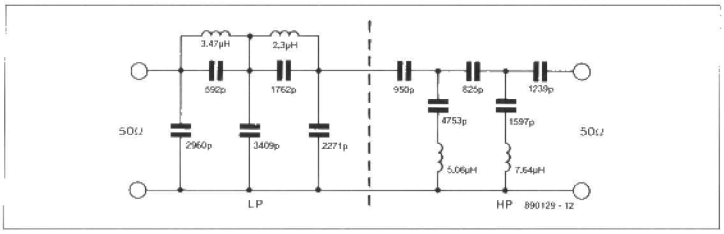



Fig. 2. Band-pass filter for the top band. The –1 dB edges are at 1.8 MHz and 2.0 MHz. The pass band ripple is 1 dB. The –40 dB points are at 1.479 MHz and 2.434 MHz. The frequencies of infinite attenuation are at: LF 1.025 MHz and 1.44 MHz; HF 2.5 MHz and 3.51 MHz.

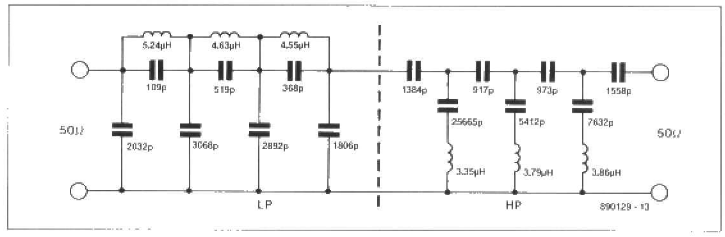



Fig. 3. Band-pass filter for the top band. Attenuation at 1.8 MHz and 2.0 MHz is 0.18 dB. The pass band ripple is 0.18 dB. The –80 dB points are at 3.178 MHz and 1.132 MHz. The frequencies of infinite attenuation are at: LF 0.9269 MHz, 1.1106 MHz and 0.54202 MHz; HF 3.88 MHz, 3.241 MHz and 6.641 MHz

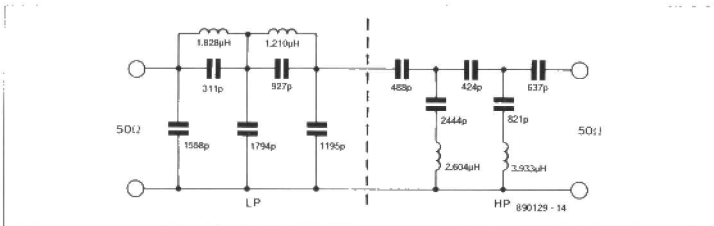



Fig. 4. Band-pass filter for the 80 m band. The –1 dB edges are at 3.5 MHz and 3.8 MHz. The pass band ripple is 1 dB. The –40 dB points are at 2.875 MHz and 4.624 MHz. The frequencies of infinite attenuation are at: LF 2.8 MHz and 1.994 MHz; HF 4.75 MHz and 6.669 MHz.

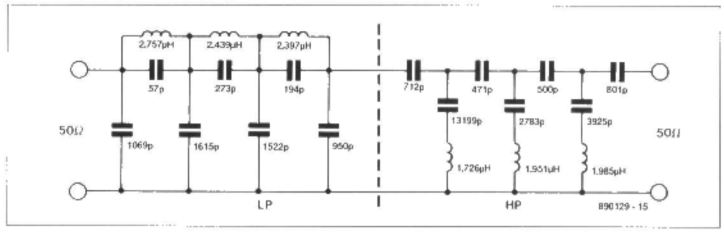



Fig. 5. Band-pass filter for the 80 m band. Attenuation at 3.5 MHz and 3.8 MHz is 0.18 dB. Pass band ripple is 0.18 dB. The –80 dB points are at 2.202 MHz and 6.038 MHz. The frequencies of infinite attenuation ar at: LF 1.053 MHz, 1.802 MHz and 2.159 MHz; HF 6.158 MHz, 7.378 MHz and 12.619 MHz.

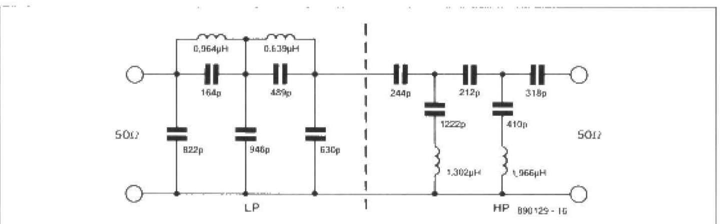



Fig. 6. Band-pass filter for the 40 m band. The –1 dB edges are at 7.0 MHz and 7.2 MHz. The pass band ripple is 1 dB. The –40 dB points are at 5.751 MHz and 8.762 MHz. The frequencies of infinite attenuation are at: LF 3.988 MHz and 5.6 MHz; HF 9.0 MHz and 12.63 MHz.

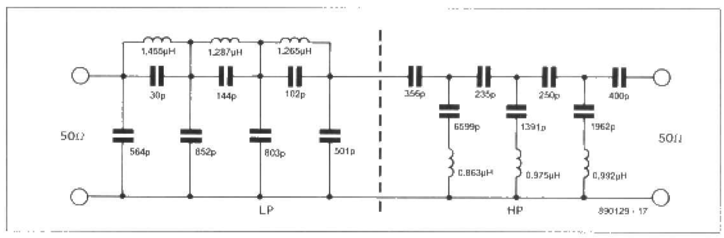



Fig. 7. Band-pass filter for the 40 m band. Attenuation at 7.0 MHz and 7.2 MHz is 0.18 dB. The pass band ripple is 0.18 dB. The –80 dB points are at 4.405 MHz and 11.44 MHz. The frequencies of infinite attenuation are at: LF 2.107 MHz, 3.604 MHz and 4.319 MHz; HF 11.668 MHz, 13.981 MHz and 23.91 MHz.

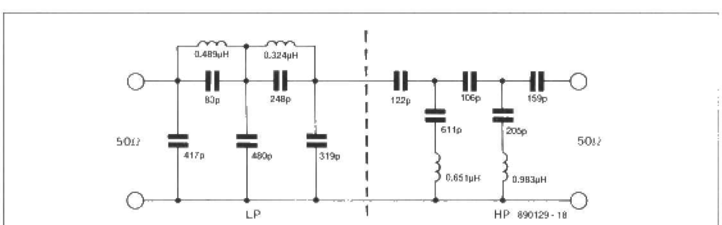



Fig. 8. Band-pass filter for the 20 m band. The –1 dB edges are at 14.0 MHz and 14.2 MHz. The pass band ripple is 1 dB. The –40 dB points are at 11.50 MHz and 17.28 MHz. The frequencies of infinite attenuation are at: LF 7.977 MHz and 11.2 MHz; HF 17.75 MHz and 24.92 MHz.

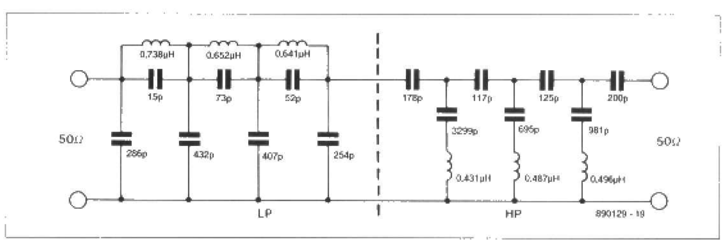



Fig. 9. Band-pass filter for the 20 m band. Attenuation at 14.0 MHz and 14.2 MHz is 0.18 dB. The pass band ripple is 0.18 dB. The –80 dB points are at 8.810 MHz and 22.564 MHz. The frequencies of infinite attenuation are at: LF 4.215 MHz, 7.209 MHz and 8.638 MHz; HF 23.01 MHz, 27.57 MHz and 47.156 MHz.

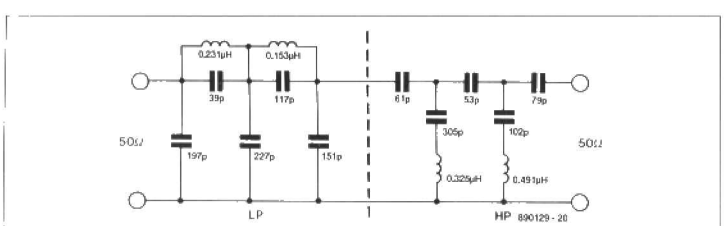



Fig. 10. Band-pass filter for the 10 m band. The –1 dB edges are at 28 MHz and 30 MHz. The pass band ripple is 1 dB. The –40 dB points are at 23 MHz and 36.51 MHz. The frequencies of infinite attenuation are at: LF 15.954 MHz and 22.4 MHz; HF 37.5 MHz and 52.65 MHz.

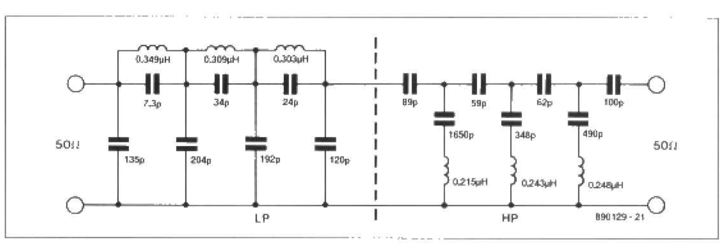



Fig. 11. Bandpass filter for the 10 m band. Attenuation at 28 MHz and 30 MHz is 0.18 dB. The pass band ripple is 0.18 dB. The –80 dB points are at 17.62 MHz and 47.6 MHz. The frequencies of infinite attenuation are: LF 8.43 MHz, 14.419 MHz and 17.277 MHz; HF 48.619 MHz, 58.254 MHz and 99.625 MHz.

# BUDGET FM RECEIVER

J.Bareford

**Most receivers for 30 MHz and up contain a fair number of semiconductors. Although super-regenerative receivers have been built with few transistors, their performance is generally rather poor. The present four-transistor receiver is of the super-heterodyne type, and demonstrates the feasibility of building a sensitive FM receiver from a minimum number of components.**

Although the super-heterodyne receiver requires relatively many sub-circuits, it is probably the best choice when a receiver for the VHF and UHF ranges is to be built.

The first building block of the 'superhet' is the input stage that filters and raises the aerial signal to a level suitable for the second block, the mixer. The filters are required to ensure sufficient image rejection. The third block is the local oscillator. This circuit is tuneable and drives the other input of the mixer. The output of the mixer is connected to the input of the intermediate frequency (IF) amplifier that provides the required selectivity. The IF amplifier is generally dimensioned for relatively high gain and a bandwidth not greater than strictly required. The last building block of the superhet is, in principle, the demodulator.

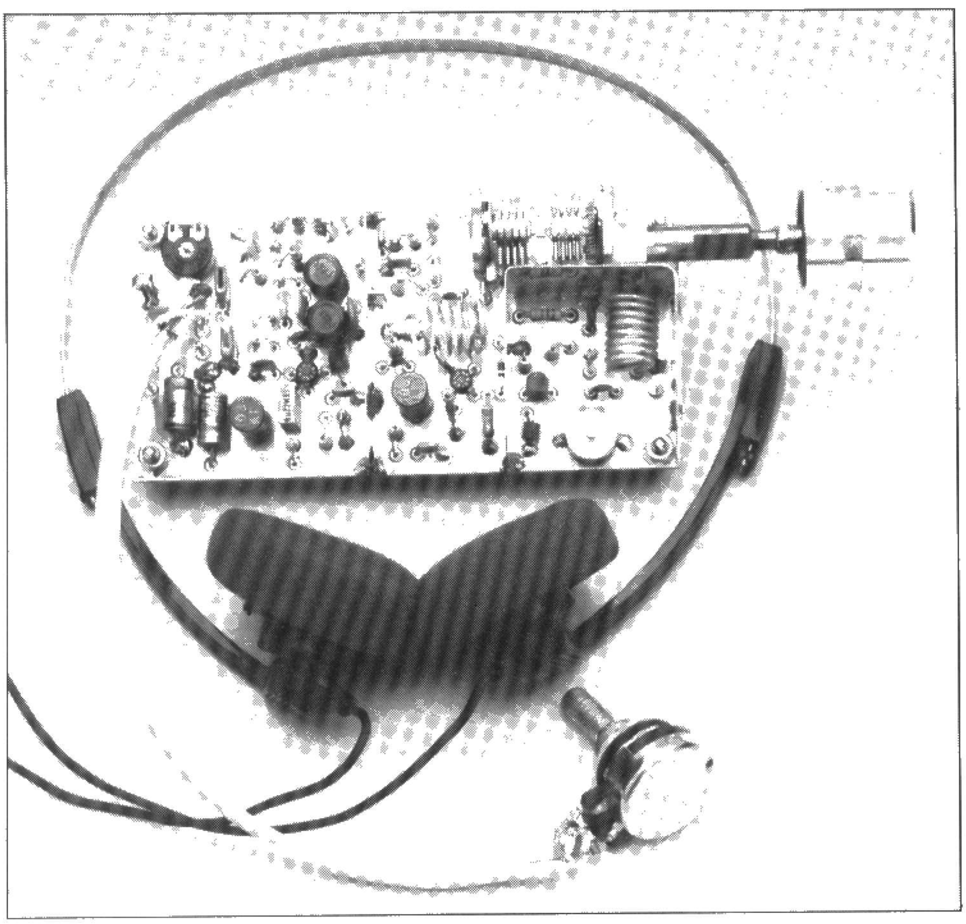
In most receivers of the super-heterodyne type, each of the above blocks uses at least one transistor — the IF amplifier usually requires several. Less conventional receiver design, however, allows the number of transistors to be reduced drastically.

## Active down, passive up

The circuit diagram of Fig. 1 shows that the number of active components has been kept to a minimum, while relatively many passive parts are used. Only three transistors take care of all the high-frequency functions, and one is used as an audio amplifier.

Transistor $T_1$ and associated parts form the RF input amplifier that raises the aerial signal. A grounded-base circuit is used to achieve acceptable matching of the input cable impedance. Since it is fairly heavily damped, tuned circuit $L_1$-$C_1$ has a relatively low loaded Q (quality) factor. The tuned circuit can not, therefore, ensure the receiver's selectivity, but this is not a problem because the intermediate frequency is low at 200 kHz, making adequate image rejection impossible anyway. The upshot of this is that two frequencies, spaced at 400 kHz, are actually received at the same time. In practice, this means that all stations simply occur twice when the receiver is tuned across the VHF FM broadcast band.

The use of a pnp transistor in the RF input stage allows the amplified signal at the collector to be coupled direct to gate-1 (g1) of MOSFET $T_2$.



## Self-oscillating mixer

The self-oscillating mixer is, admittedly, infamous with many high-frequency enthusiasts. The design presented here, however, suffers none of the disadvantages like poor mixing characteristics, noise, and 'pulling' of the input stage, associated with traditional bipolar transistor designs. MOSFET $T_2$ is basically used in a standard mixer configuration because the oscillator signal is available at gate-2, and the received signal at gate-1. Unconventionally, however, the mixer has a feedback network between the source and gate-2. This is achieved with parts $L_3$-$C_5$ and $D_1$ that cause the mixer to oscillate. Varicap diode $D_1$ has no effect on the receiver tuning (this is taken care of by $C_5$), but serves to rectify the RF voltage across $L_3$. The rectified voltage makes gate-2 of the MOSFET a little positive with respect to ground. The internal gate-source capacitance serves to buffer this d.c. level, which raises the transconductance of the MOSFET for the signal applied to gate-1, but at the same time lowers the transconductance for the signal applied to gate-2. The combined regulation causes the overall amplification in the oscillator to stabilize at a certain level, and results in a relatively clean output signal.

High-frequency components in the mixer output signal are suppressed by $C_6$ and series network $C_8$-$L_5$, so that a clean signal is applied to the IF amplifier, $T_3$. Choke $L_4$ prevents the IF signal being

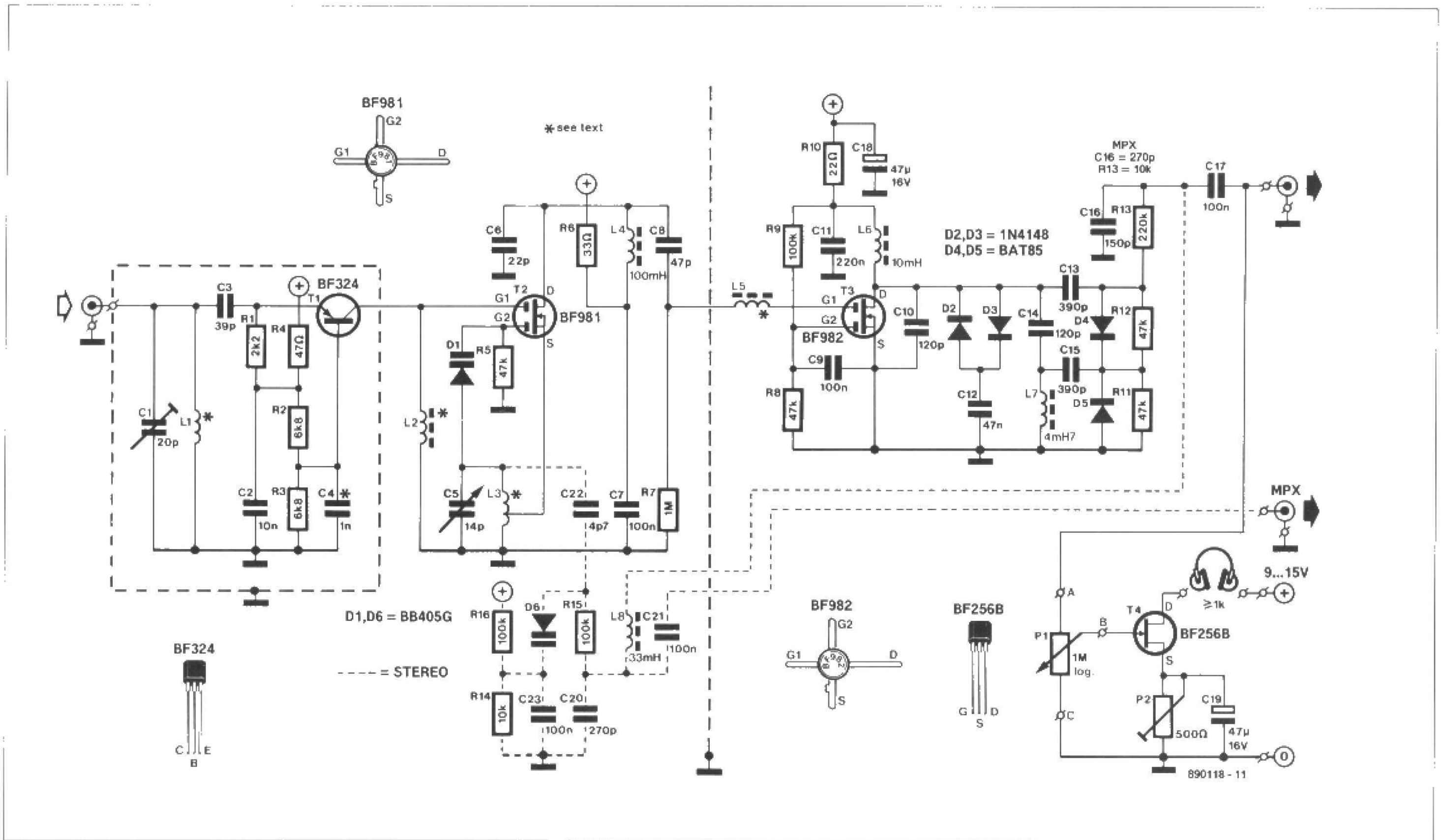



**Fig. 1. Circuit diagram of the budget FM receiver. The dashed parts are for the optional AFC extension, which also allows a stereo demodulator to be connected to the MPX output..**

short-circuited by the decoupled positive supply rail.

## One-transistor IF amplifier

The fairly high conversion gain of mixer/oscillator $T_2$ allows a single transistor to be used as the IF amplifier. A dual-gate MOSFET is used as in the mixer/oscillator, but $T_3$ is a Type BF982 rather than a BF981 because a higher transconductance is required. The function of $L_6$ is similar to that of $L_4$.

The IF amplifier is not tuned — the only IF filter in the receiver, $L_7$-$C_{14}$, is located behind the limiter, $D_2$-$D_3$.

## Discriminator

Apart from its function as the only IF filter in the FM receiver, tuned circuit $L_7$-$C_{14}$ acts as an FM demodulator. When the series combination resonates at about 200 kHz, the voltage across the inductor, $L_7$, has the same amplitude as that across $C_{14}$, but a phase difference of 180°. For frequencies above the resonance frequency, the reactance of the inductor is greater than that of the capacitor. This means that the voltage on the inductor is greater than that across the capacitor. For frequencies below the resonance frequency, the capacitor voltage exceeds the inductor voltage, although the phase difference is still 180°.

Since the IF signal is frequency modulated, the voltages across $L_7$ and $C_{14}$ can be rectified to recover the modulation signal. Diode $D_4$ rectifies the voltage across $C_{14}$, and $D_5$ the voltage across $L_7$. Since the cathodes are interconnected, the total direct voltage with respect to ground equals the difference between the individual diode voltages. This difference is, of course, 0 V at the resonance frequency of $L_7$-$C_{14}$, while it is positive or negative depending on the instantaneous signal frequency, which is a function of the modulation signal because frequency-modulated (FM) signals are received.

The currents supplied by $D_4$ and $D_5$ during their conductive half-cycles are used to charge $C_{13}$ and $C_{15}$ respectively. These capacitors are slowly discharged by parallel resistors $R_{12}$ and $R_{11}$.

## Headphone amplifier

The demodulated audio signal is applied to low-pass filter $R_{13}$-$C_{16}$ which attenuates frequencies above 5 kHz and so forms a basic de-emphasis.

The filtered AF signal is fed to headphone amplifier $T_4$. This FET provides sufficient power only if headphones with an impedance greater than 1 kΩ are connected. For higher output volume and/or a lower output impedance, use an LM386-based AF amplifier.

## AFC and stereo

Automatic frequency control (AFC) is fairly simple to implement in the receiver by adding the parts connected by the dashed lines in the circuit diagram. The added parts create a feedback path between the audio signal and tuning element $D_6$. The result is automatic frequency correction, so that the receiver remains tuned to a station in spite of frequency drift of the local oscillator owing to temperature changes. An additional benefit of the feedback circuit is an improvement in the linearity of the audio signal, which results in a better high-frequency response.

The high boost achieved by the addition of the AFC parts allows a multiplex (MPX) demodulator to be connected to output MPX to enable stereo reception. In a stereo MPX signal, the L–R difference signal is modulated on to a subcarrier at 38 kHz. Without the AFC circuit, this signal would be difficult to recover from the normal AF output because of the roll-off at 5 kHz.

The use of a stereo FM demodulator, which is not discussed here, also requires $C_{16}$ and $R_{13}$ to be changed to the values indicated in the circuit diagram.

## Construction

The printed circuit board shown in Fig. 2 is double-sided, but not through-plated. The largely unetched component side functions as a ground plane that prevents parasitic coupling between components and PCB tracks.

Drill the PCB and sort the components before fitting them. A large number of parts is mounted vertically. As the component density is fairly high, these parts must be installed carefully, and soldered accurately to prevent short-circuits. Solder component terminals at both PCB sides if the solder spot at the track side is not connected to a track or other spot(s).

The mounting of SMD capacitor $C_4$ at

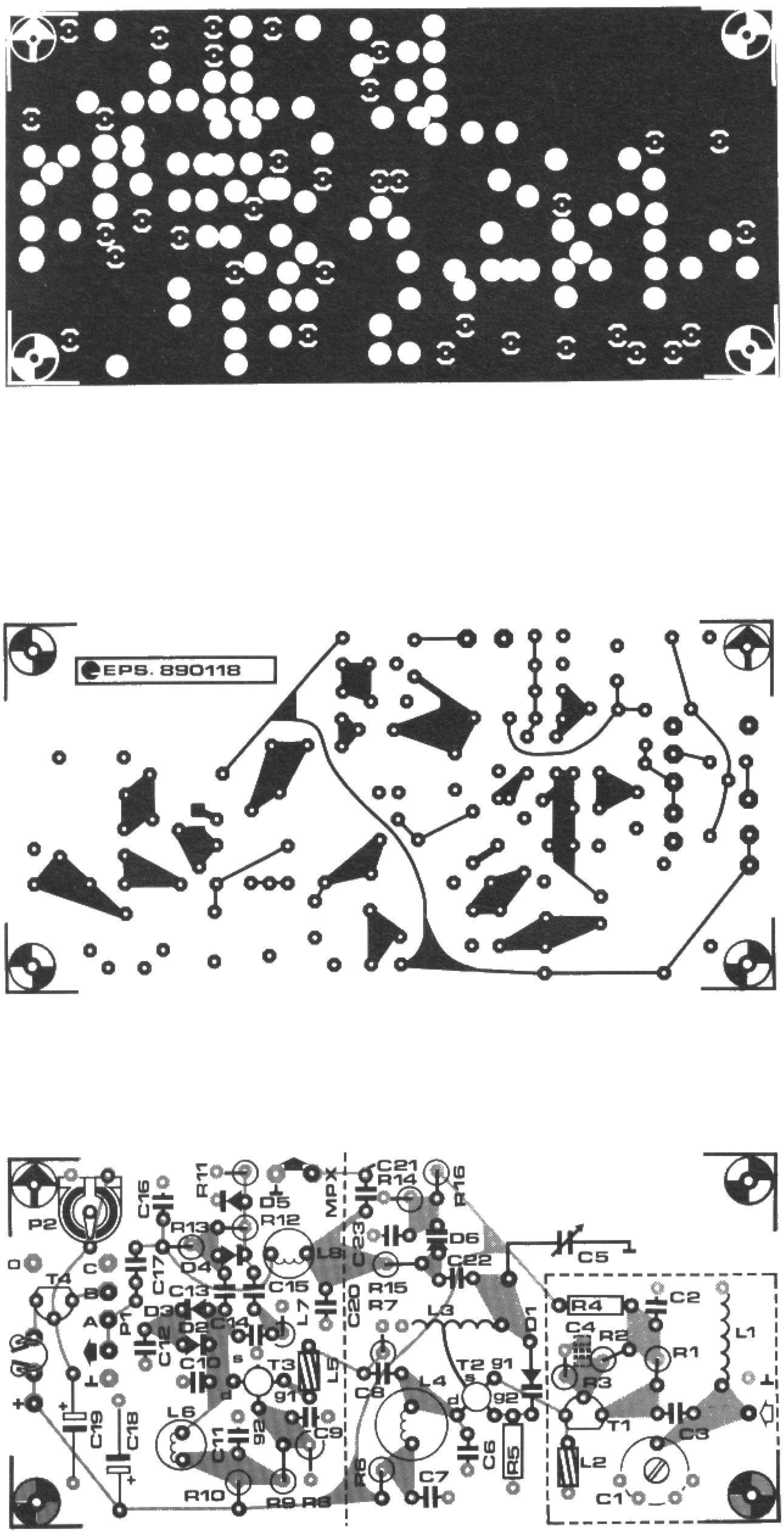



Fig. 2. Double-sided PCB for the FM receiver. Fit screening plates over the dashed lines.

**Parts list**

**Resistors:**

$R_1$ = 2k2
$R_2$;$R_3$ = 6k8
$R_4$ = 47Ω
$R_5$;$R_8$;$R_{11}$;$R_{12}$ = 47k
$R_6$ = 33Ω
$R_7$ = 1M0
$R_9$;$R_{15}$*;$R_{16}$* = 100k
$R_{10}$ = 22Ω
$R_{13}$ = 220k (10k*)
$R_{14}$* = 10k
$P_1$ = 1M0 logarithmic potentiometer
$P_2$ = 500Ω preset H

**Capacitors:**

$C_1$ = 20p foil trimmer (green)
$C_2$ = 10n ceramic
$C_3$ = 39p
$C_4$ = 1n0 SMD
$C_5$ = 14p tuning capacitor
$C_6$ = 22p
$C_7$;$C_9$;$C_{17}$;$C_{21}$*;$C_{23}$* = 100n ceramic
$C_8$ = 47p
$C_{10}$;$C_{14}$ = 120p
$C_{11}$ = 220n
$C_{12}$ = 47n
$C_{13}$;$C_{15}$ = 390p
$C_{16}$ = 150p (270p*)
$C_{18}$;$C_{19}$ = 47μ; 16 V
$C_{20}$* = 270p
$C_{22}$* = 4p7

**Semiconductors:**

$D_1$;$D_6$* = BB405G
$D_2$;$D_3$ = 1N4148
$D_4$;$D_5$ = BAT85
$T_1$ = BF324
$T_2$ = BF981
$T_3$ = BF982
$T_4$ = BF256B

**Inductors:**

$L_1$;$L_2$;$L_3$;$L_5$ = home-made inductor; winding details are given in the text. Parts required: 1 mm dia. silver-plated wire and 2 ferrite beads (3mm).
$L_4$ = 100 mH radial choke with ferrite encapsulation, e.g., Toko 181LY-104 (Cirkit).
$L_6$ = 10 mH radial choke with ferrite encapsulation, e.g., Toko 181LY-103 (Cirkit).
$L_7$ = 4mH7 axial choke, e.g. Siemens B78108 (Cirkit stock no. 35-71475).
$L_8$* = 33 mH radial choke with ferrite encapsulation, e.g., Toko 181LY-333 (Cirkit).

**Miscellaneous:**

PCB Type 890118 (not available through the Readers Services).
High-impedance headphones or earpiece.
Co-axial aerial socket: e.g., BNC or SO-239.
Metal enclosure.
Mains adapter socket.

* AFC option

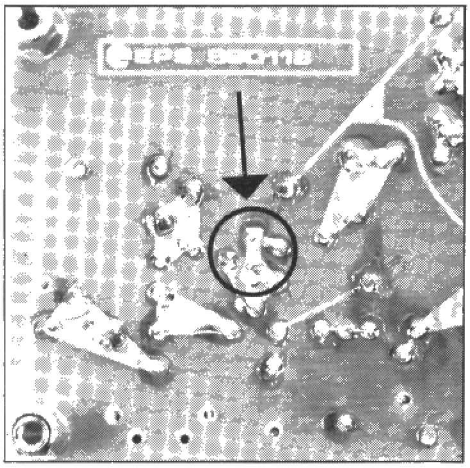


**Fig. 3. SMD capacitor C4 is soldered at the track side of the board, between the spots provided for R3.**

the track side of the PCB is shown in the photograph of Fig. 3.

The source terminal of $T_2$ is connected direct, via a short wire, to the tap on $L_3$. The other terminals of the MOSFET are inserted in the appropriate PCB holes.

The tuning capacitor used in the prototype was a 14pF single-gang type from Schwaiger. Other types with a different mechanical construction are also suitable, provided the capacitance range is about 5pF to 14pF. The Schwaiger capacitor must be modified before it can be mounted on to the PCB: cut off the terminals, and solder a single wire at the other side of the capacitor body. Solder quickly to prevent the stator package being dislodged and causing short-circuits with the rotor blades. Scratch off the coating from the sides of the capacitor body so that the device can be soldered on to the PCB.

Install 15 mm high tin plate or brass screens on to the PCB. The placement is indicated by the dashed lines on the component overlay.

The receiver must be powered from a low-noise, regulated 12V source. A simple mains adapter will generally be unsuitable unless additional regulation is installed. In most cases, a Type 7812 regulator with associated decoupling capacitors is adequate for this purpose, but make sure that the unregulated input voltage is between 15 and about 20 V.

## Inductors

Wind four inductors as detailed below.

**Inductor $L_1$:** close-wind 8 turns of 1 mm dia. (SWG20) silver-plated wire on to an 8 mm drill. Make sure that the ends of the inductor are in the same plane (see Fig. 4.). Then bend the ends at right angles, keeping them central under the turns as indicated by the dashed lines in Fig. 4. Now space the turns evenly so that the terminal wires align with the holes in the PCB.

**Inductor $L_3$:** this is made as $L_1$, but with 6 turns rather than 8. The location of the tap is shown in Fig. 4.

**Inductor $L_2$:** wind 2 turns of 0.2 mm dia. (SWG36) enamelled copper wire through a 3 mm long ferrite bead (see Fig. 5a).

Inductor $L_5$: as $L_2$, but with 4 turns (see Fig. 5b).

The remaining inductors in the receiver are ready-made chokes. Suggested types are given in the Parts List.

## Alignment

Apply 12 V d.c. to the completed receiver, connect an aerial (75 Ω) and headphones. Set the tuning capacitor to maximum capacitance and compress or stretch $L_3$ until the local oscillator operates at 88 MHz. Check this with a frequency meter or another FM receiver. If necessary, reduce the number of turns of $L_3$.

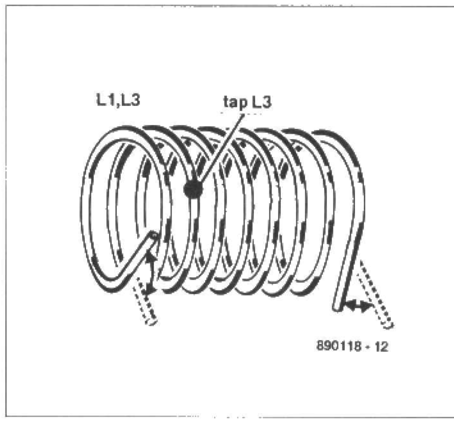



**Fig. 4. Winding details of inductors $L_1$ and $L_3$.**

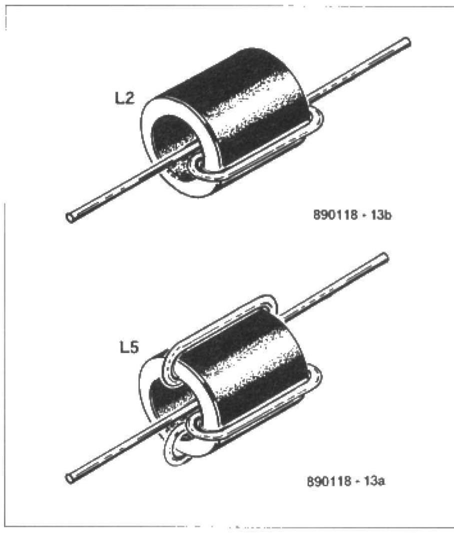



**Fig. 5. Winding details of inductors $L_2$ and $L_5$.**

Tune to a weak transmission and peak $C_1$ for maximum output volume. Finally, adjust preset $P_2$ for an acceptable compromise between distortion and output volume.

# ELECTRONICS SCENE

### British Gyros Will Line Up New Intelsat Comsats

Gyros from Ferranti will ensure the initial line up and correct attitude on station of the new series of American Intelsat-7 communications satellites.

Similar Ferranti equipment has already been used successfully in satellites such as IRAS, Exosat, X4 (Miranda) and for the Spacelab Instrument Pointing System (IPS). Moreover, it has been provided for the Rosat X-ray Astronomical Satellite and the Olympus communications satellite

### Analogue Touch Sensors

Analogue touch sensors that use surface chemistry and surface electronics for input via visual display units (VDUS) have been developed by John McGavin & Co.

The sensors consist of two conductive coatings applied to a substrate of polyester or polycarbonate.The faces are separated by clear dielectric spacer dots and are brought into electrical contact only when actuated by the pressure of a finger.

### Computer Graphics Metafile

A working demonstration of computer graphics metafile (CGM), the British and ANSI standard that provides integration of different computer graphics products and the exchange of information and images between systems, will be given at the Computer Graphics Exhibition and Conference from 7 to 9 November at the Alexandra Palace, London.

### Simulator for Satellite Signals

A simulator that does a job similar to those used to train aircraft pilots has been developed by STC to check on the accuracy of Global Positioning Systems (GPS) used worldwide for navigating both civil and military craft on land, sea and in the air. The company says that its STR2700 simulator will contribute to still more accurate navigation from satellite signals.

Global Positioning Systems relay on signals from 18–21 special navigational satellites in orbit around the earth, of which the average user can 'see' up to five at any given moment.

# AMATEUR COMMUNICATION RECEIVERS: STILL A CHALLENGE?

by A. B. Bradshaw

## The valve era

Over the past forty years or so, communication receiver design has undergone quite a revolution. In the days before transistors and ICs, there were some remarkably good receivers around. Typical among these were the AR88, the Hamerlund Super-Pro, the Marconi CR100, the BC348, and the Racal RA17.

After the end of the Second World War, many radio amateurs were using either one of these classical designs or one of the many ex-military receivers that had come on to the surplus market.

A large number of amateurs showed great ingenuity in the use of various items of military equipment to make up their station receiver, and sometimes their transmitter as well. There was a lot of ex-services expertise about and a considerable amount of technical discussion seemed to take place over the airwaves. Cobbling together all this readily obtainable gear was not entirely caused by the non-availability of proprietary amateur equipment: most of us being broke had something to do with it as well!

Towards the end of the valve era, there occurred a number of technical developments in radio valve technology that had a direct bearing on communication receiver design. One of these was the appearance of the frame grid pentode, like the E183.

These new valves, 465 kHz IF transformers with a good Q, and ex-government quartz crystals, such as the FT241/243, helped to achieve respectable IF response shapes for reception of the increasingly popular single-sideband (SSB) transmissions.

At the same time, wide-range, stable automatic gain control (AGC) was becoming the norm, its control voltage no longer derived from the incoming carrier.

The emergence of the long-life stable double triodes, like the E88CC, originally developed for the then embryonic computer industry, further helped to improve amateur communication receiver design.

Another milestone was the introduction of the beam deflection mixer valve, like the 6AR8 and the 7360, which were developed for the American colour TV market. The remarkably linear mixing and large-signal handling capabilities of these new valves soon caught the eye of receiver designers and it did not take long before manufacturers like Collins, Squires-Sanders, Drake, and so on were incorporating them in their new receivers.

At about this time, a superb receiver, the Thornley G2DAF design, appeared on the UK amateur scene. Many of these excellent receivers were built and had a profound influence on our thinking of what kind of performance could be achieved with the technology then available. I built my own and well remember the pleasure of using the receiver, which had the knife-edge selectivity of the Kokusai mechanical filter Type MF455-10.K.

By then, we had the ingredients necessary for meeting the specification for a good communication receiver:

- good IF shape factor (in spite of the low IF resulting from the multi-conversion necessary for the HF end);
- stable conversion oscillators, necessary for the increasingly popular SSB mode of transmission;
- ease of tuning with mechanical S/M drives (Eddystone 898, and so on).

Nevertheless, these receivers still had some serious short-comings. They were complex (at the time); they usually embodied lots of ganged switching of tuned circuits; they used relatively expensive wound components; their front-end alignment and tracking, particularly in general coverage designs, was difficult; and lastly, these 'magificent machines' could certainly not be regarded as portable.

## The solid-state era

The transition to solid state electronics was not a sudden occurrence, and for some years hybrid designs were very popular in the amateur press. Although these designs still used valves in their front-ends, much of the remaining circuitry had become solid-state. These early solid-state devices, however, could not produce the good intermodulation and cross modulation performance of their valved predecessors.

Over the past decade, solid-state devices have improved enormously, however, and present-day communication receivers have very real benefits compared with those of yesteryear.

Unfortunately, in my view, we have allowed the Japanese industry to dominate the manufacture and design of good-quality communication receivers. This is particularly disappointing in view of our own earlier performance. In the solid-state era we have managed to produce some innovative designs, but they are few and far between.

Nevertheless, the radio amateur remains in a unique position. The receiver manufacturer is hamstrung by severe economic restraints and market forces. The amateur designer and constructor, on the other hand, is still at liberty to explore and indulge his fancy in ways that would be out of the question for the professional designer. I am not suggesting for one moment that the radio amateur can challenge the Japanese giants. Nevertheless, there is still much innovation in Britain, well documented in a variety of books, technical articles, application notes, and so on.

## Modern home construction

If we regard the modern communication receiver at a system level, we have a good opportunity to see what some British manufacturers and suppliers have on offer

**RF amplifiers:** Plessey Types SL600; SL611C; SL612; SL1610C; SL1611C; SL1612C.

**High performance mixers:** Plessey Type SL6440A/C (+30 dBm intercept point); Siliconix Type Si8901 double-balanced mixer (+35 dBm intercept point); various diode bridge ring devices, from the MD108 up to the SRA3 (£28 from Cirkit).

**IF shaping filters**: ceramic and mechanical filters are available for the lower IFs (455 kHz), while for the higher IFs there are quartz crystal lattices up to 10.7 MHz are available in bandwidths suitable for AM, SSB and CW from Cirkit.

**IF amplifiers**: three Plessey Type SL612 ICs will give most of the gain required in a normal IF amplifier.

**Demodulators for AM, SSB, and CW**: Plessey Types SL6700A and SL624.

**AGC generators**: rather a limited choice

here, but the Plessey SL620 and SL621C ICs are well proven.

This list shows that there is a good home-bred range of building blocks, although I still feel that there are areas of design that have been neglected. Some of these are receiver front-end filtering for the amateur bands, local oscillator design (either upper conversion synthesis or limited-range BFO conversion systems), while the required noise floor specification for VHF synthesizers is a real challenge.

## Conclusion

As I glance through yesteryear's copies of *RSGB Bulletin*, *RAD COM*, and others, I can not but be struck by the falling off in interest in innovative design. Can this malaise be halted? I certainly hope so. What are you going to do about it?

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"Receiver Noise Figure, Sensitivity, and Dynamic Range: What the Numbers Mean" by J. Fisk, *Ham Radio*, Oct. 1975.
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"The RX80 Mk2" by A.L. Bailey, *Rad Com*, Jan. 1981 (in 6 parts).
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"A Home Built Frequency Synthesizer for 45–75 MHz" by J. Crawley, *Rad Com*, August 1986 (in 2 parts).

# ELECTRONICS SCENE

### Encryption System May Be Basis for International Standard

A high security encryption system developed by GEC Plessey Telecommunications (GPT) for video conferencing, may provide the basis for a new international standard.

Called B-Crypt, the system is selling all over the world, demonstrating a strong international demand for a dedicated encryption system.

B-Crypt operates on two 56-bit security codes. The first is a crypto-variable key that can be calculated automatically by the system or input manually. The second is an initialization vector. This is a random number that is changed and transmitted to the receiver every 32 milliseconds. The actual encryption key, a factor of these two security codes, is never transmitted over the link and is, therefore, irretrievable by a third, unauthorized party. While encryption is active, all timeslots carrying video, audio and data are fully encrypted. Timeslots zero and two are unencrypted at all times for network signalling purposes.

Control of both the internal and standalone systems is by a terminal via a standard RS232C serial port fitted to the rear panel. Under normal circumstances, initialization of encryption will occur when two suitably equipped locations of a video teleconference request the transmission to be encrypted via their local control terminal. If only one location makes the request, encryption will not occur and a warning message will be displayed at the control terminal. Both codec and standalone systems are fitted with 'Force Encryption' capability, which enables them to react automatically to an encryption request from a remote terminal with local intervention.

### Another Step towards Global ISDN

British Telecom has taken another step towards a global integrated services digital network (ISDN) with the introduction of new digital switched services linking the UK, Japan and the United States.

Customers of British Telecom, KDD and AT&T who have compatible equipment will now be able to use advanced applications such as new generation (Group 4) facsimile, and the exchange of data files at high speed.

The initial service will provide a circuit switched data capability operating at 56 kbit/s or 64 kbit/s. In time, this will be enhanced to provide supplementary services such as calling line identity and closed user groups and functions such as terminal identification.

British Telecom's ISDN, named integrated digital access—IDA—is available either as single-line or as multi-line.

### Anglo-French-US Partners in Global Network

British Telecom, AT&T and France Telecom have signed a contract to provide the first phase of a communications network that will eventually span six continents.

By the end of this year, the first stage of the project will link the three hubs of London, New York and Paris, thus extending the General Electric voice and data network from the US into major European locations. British Telecom, from its network management centre in London, will be the hub for the UK, the Netherlands, Ireland and Spain.

Other European locations will be linked next year, and phase 2 is planned to link the Far East, Middle East and South America. The global network will provide GE with voice, data and videoconferencing services.

### GPT Credit Card Payphones for Federal Germany

Federal Germany's first public credit card payphone service, which uses British equipment, has recently gone into service at Frankfurt Airport. Others have been installed at the airports of Dusseldorf, Hamburg and Munich.

The payphone equipment was developed and supplied by GEC Plessey Telecommunications (GPT) to the German Bundespost. GPT has now supplied the equipment to 58 customers in 44 countries worldwide.

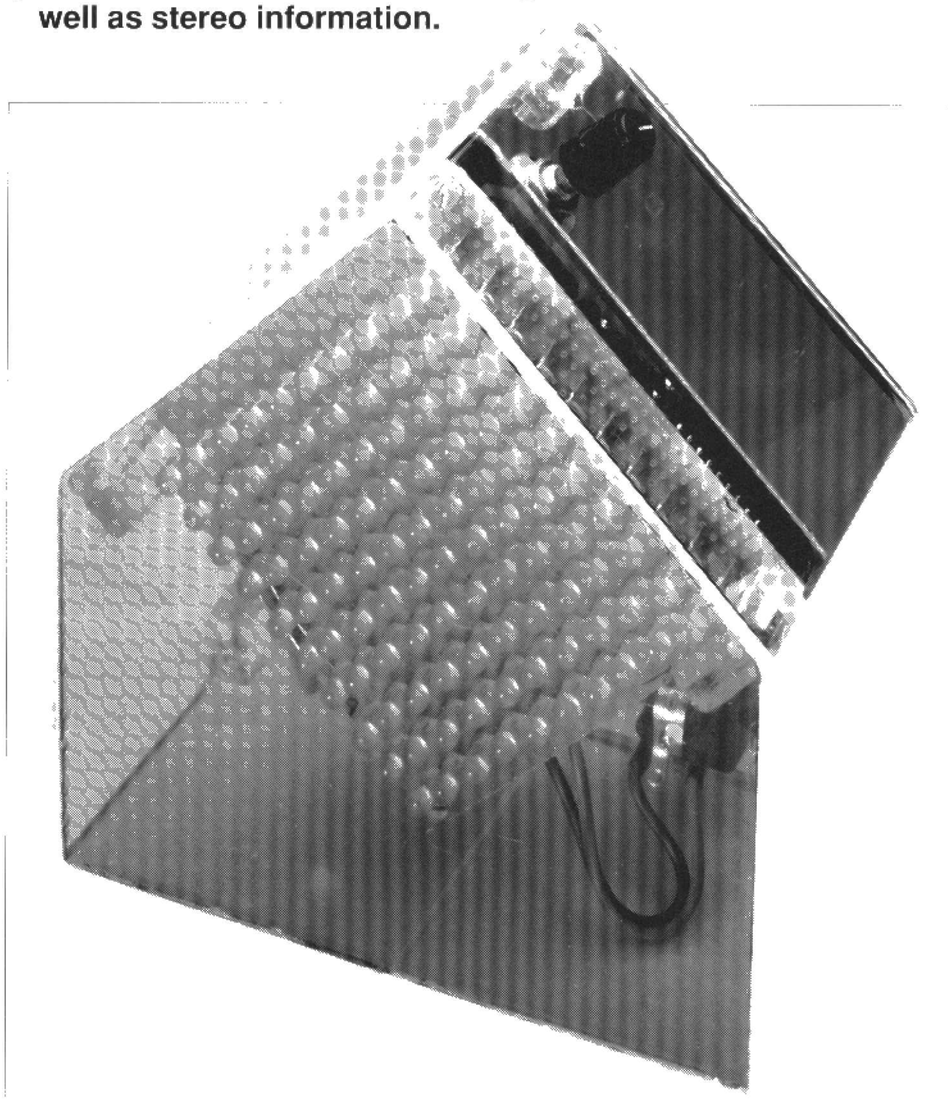
The payphones accept payment by standard international credit cards, such as Amex, Access, Diners and VISA.

Voice messages on the system are used to assist the caller, if necessary, and are provided with dual language capability, normally English plus the language of the country involved.

# STEREO VIEWER

**C.J. Ruissen & A.C. van Houwelingen**

**This electronic ornament is basically an unconventional VU-meter. A square matrix composed of 10×10 LEDs indicates signal volume as well as stereo information.**



The circuit is perhaps best qualified as a simple X-Y display for audio signals, and the displayed patterns are, therefore, not unlike Lissajous figures. The heart of the circuit is formed by an integrated circuit from National Semiconductor, the Type LM3914. At first glance, this dot/bar display driver is a quite conventional design. The IC houses ten comparators, a precision linear-scale voltage divider and a reference voltage source. The actual realization of these parts, however, gives the LM3914 a number of interesting features:

- outputs drive LEDs, LCDs, fluorescent displays or miniature bulbs
- external input selects bar or dot display mode
- simple to cascade for displays with a resolution of up to 100 steps
- internal voltage reference; adjustable between 1.2 and 12 V
- minimum supply voltage: 3 V
- current-regulated open-collector outputs
- output current programmable from 2 to 30 mA
- no multiplex switching
- input withstands ±35 V
- outputs interface direct with TTL and CMOS logic
- floating 10-step divider can be connected to a wide range of voltages, including internal reference

## Circuit description

The circuit diagram of the 10×10 LED matrix which determines the appearance of the stereo viewer is given in Fig. 2. The dimensions of the matrix result in a square arrangement. How the square is actually positioned is a matter of personal preference, and not, of course, of any circuit configuration. The introductory photograph shows the prototype which has matrix co-ordinate X1-Y1 below and X10-Y10 at the top.

The matrix arrangement allows any one of the 100 LEDs to be turned on and off individually. To select a particular LED, the relevant column, X1–X10, and row, Y1–Y10, is made high and low respectively. The circuit diagram of the row/column driver (Fig. 1) shows that two LM3914s are used: $IC_2$ forms the column driver (X-axis), and $IC_3$ the row driver (Y-axis). Both LM3914s are set to operate in the dot mode so that, strictly speaking, one row and one column are selected to light one LED at a time. The IC outputs have some overlap, however, so that two LEDs are on at the switch-over levels.

Transistors $T_2$–$T_{11}$ function as inverters. They are required because the column driver must switch to the positive supply rather than to ground. The programmable current source in $IC_3$ is set to supply the relatively small base currents for the inverter transistors. The current source in $IC_2$ is set to a much higher value to supply the required current direct to the LEDs.

The current source in the LM3914 is set in a rather unconventional manner: the output current is ten times the current supplied by the reference voltage. So, all that is required is to load the reference with a resistor. $R_{11}$ sets the output current of $IC_3$ to about 2 mA. A slightly different approach is used in the case of $IC_2$: here, an LDR (light-dependent resistor), a transistor, $T_1$, and a handful of other components form a load resistor whose value is a function of ambient light intensity. Since the output current of $IC_2$ is used for driving the LEDs, the display intensity is automatically controlled as a function of ambient light conditions. The component values used allow the LED current to vary between 8 and 25 mA.

To make sure the LEDs are completely off when they have to be off, the $\overline{LI}$ outputs of $IC_2$ and $IC_3$ are fitted with a pull-up resistor. This is required because the $\overline{LI}$ output has an auxiliary current source that is used for cascading driver chips to form a larger display. The pull-up resistors keep $T_2$ from conducting, and one

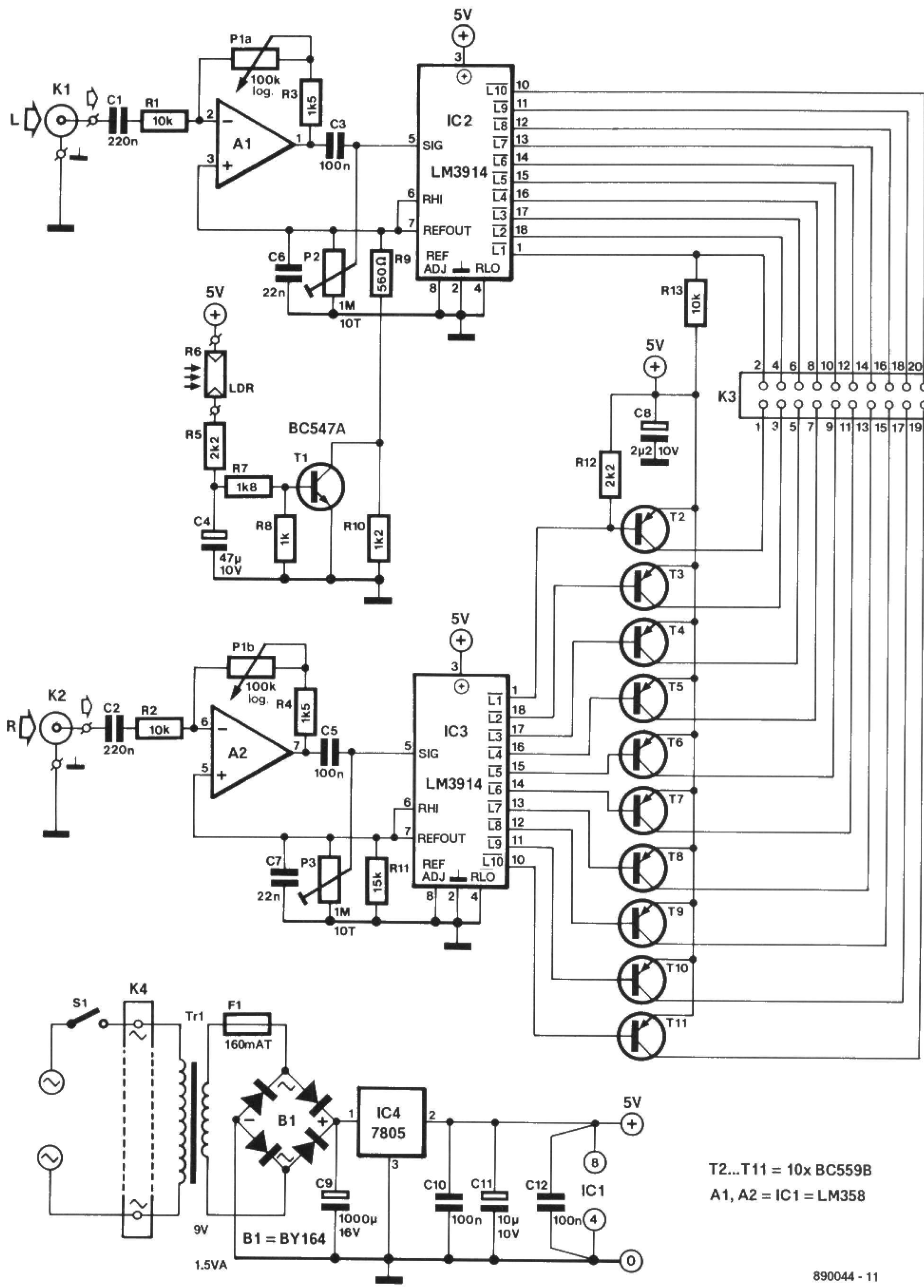



Fig. 1. Circuit diagram of the stereo viewer.

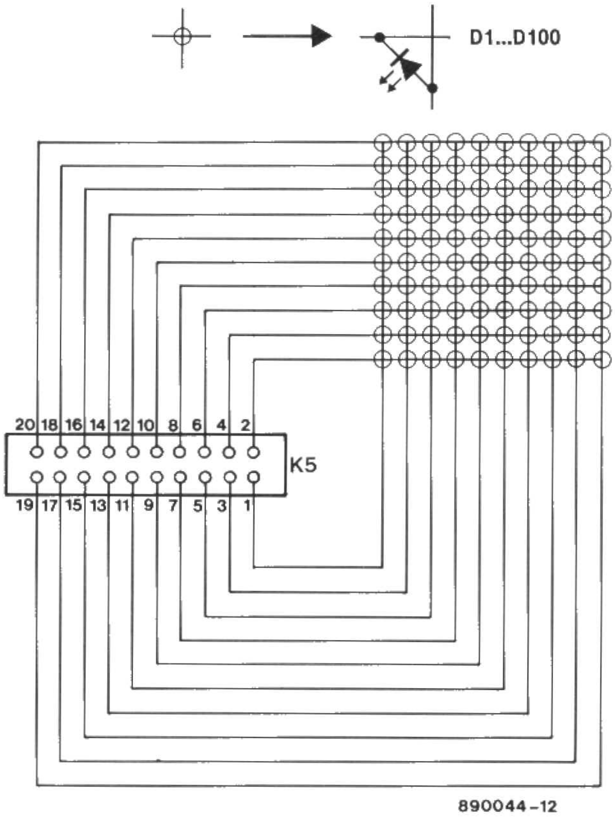



Fig. 2. Matrix configuration with 100 LEDs.

LED in the matrix from lighting, when the $\overline{LT}$ output is not actuated.

The audio signals applied to the stereo viewer are first attenuated to enable the drive levels for $IC_2$ and $IC_3$ to be set accurately. The sensitivity of the circuit is set with potentiometer $P_1$. Acceptable drive levels at the inputs are between 45 mV and 3 V.

The zero point of the matrix is shifted to the centre of the square display with the aid of bias voltages on to which the AF signals are superimposed. These voltages are obtained with multiturn presets $P_2$ and $P_3$, which are adjusted to supply half the reference voltage. A voltmeter is not required for this adjustment, because the zero indication can be seen to shift to the

**Parts list**

**Resistors (±5%):**
$R_1$;$R_2$;$R_{13}$ = 10k
$R_3$;$R_4$ = 1k5
$R_5$;$R_{12}$ = 2k2
$R_6$ = LDR
$R_7$ = 1k8
$R_8$ = 1k
$R_9$ = 560Ω
$R_{10}$ = 1k2
$R_{11}$ = 15k
$P_1$ = 100k logarithmic potentiometer; stereo
$P_2$;$P_3$ = 1MΩ multiturn preset

**Capacitors:**
$C_1$;$C_2$ =220n
$C_3$;$C_5$;$C_{10}$;$C_{12}$ = 100n
$C_4$ = 47μ; 10 V
$C_6$;$C_7$ = 22n
$C_8$ = 2μ2; 10 V
$C_9$ = 1000μ; 16 V; radial
$C_{11}$ = 10μ; 10 V

**Semiconductors:**
$B_1$ = BY164
$D_1$–$D_{100}$ = LED; dia. 5 mm
$T_1$ = BC547A
$T_2$–$T_{11}$ = BC559B
$IC_1$ = LM358
$IC_2$;$IC_3$ = LM3914
$IC_4$ = 7805

**Miscellaneous:**
$F_1$ = 160 mA fuse with PCB-mount holder.
$S_1$ = SPST mains switch.
$Tr_1$ = PCB-mount transformer 9 V; 1.5 A.
$K_1$;$K_2$ = phono socket.
$K_3$ = pin header 2×10 contacts.
$K_4$ = 2-way PCB terminal block.
$K_5$ = IDC header 2×10 contacts.
PCB Type 890044 (see Readers Services page).

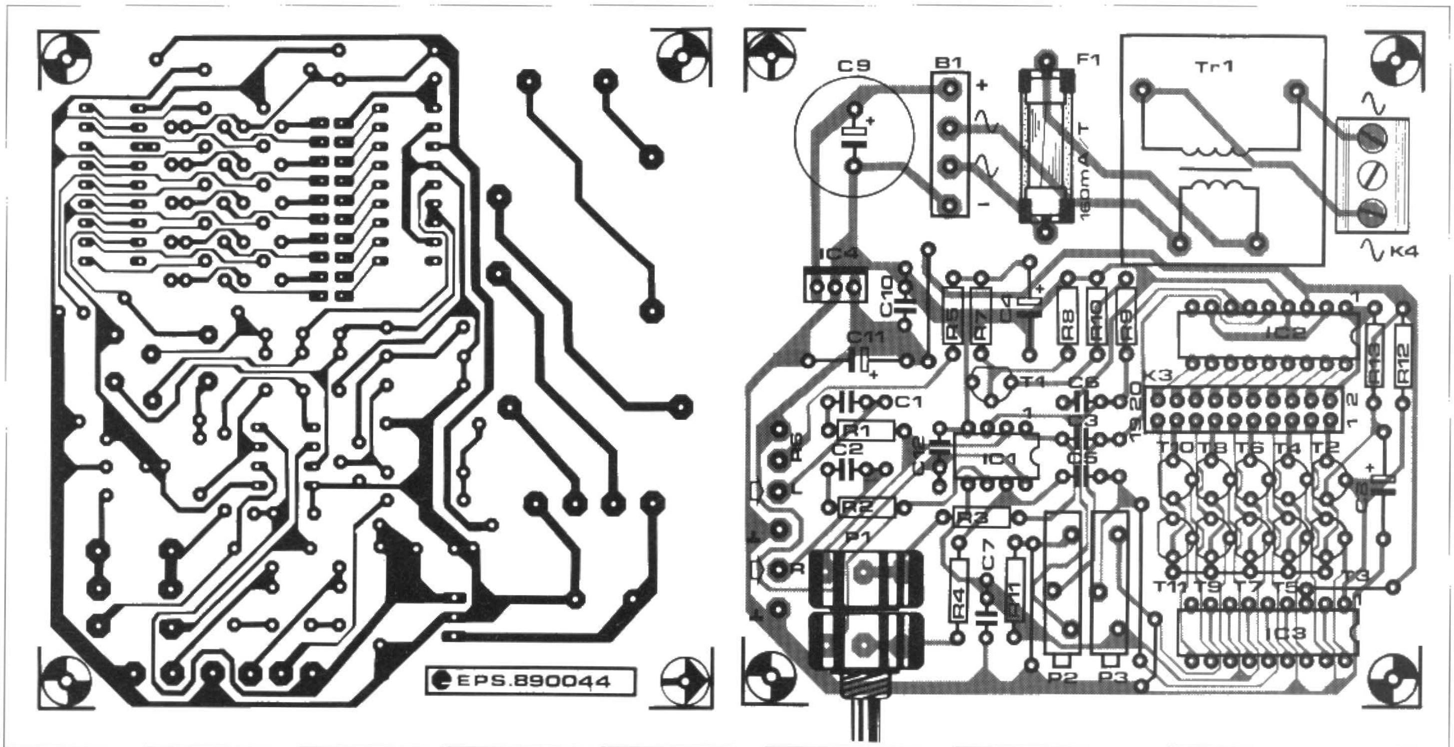



Fig. 3. Track layout and component mounting plan.

centre of the display.

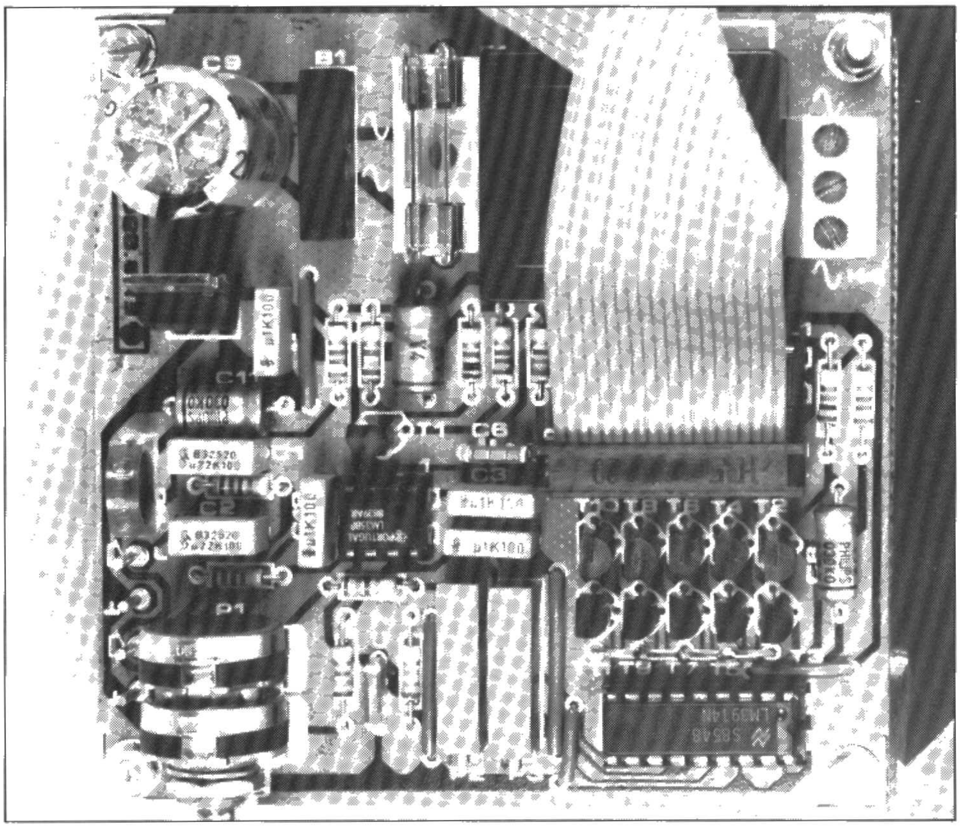
The circuit is powered by a conventional regulated 5 V supply, which is fitted on to the printed-circuit board together with the associated mains transformer.

## Construction: simple

The printed-circuit board shown in Fig. 3 accommodates all parts except the LED matrix and the LDR for the display intensity control. Populating the PCB is entirely straightforward if the wire links are installed first.

The LED matrix is built separately on a square piece of veroboard. The installation of the LEDs and the lozenge-shaped wiring at the rear of this board are greatly simplified when the matrix is turned 45° with respect to the hole pattern in the board. The LED matrix is connected via a short length of flat-ribbon cable, for which a mating 20-way pin header, K3, is provided on the main board.

The stereo viewer is simple to align: simply adjust P2 and P3 until the centre four LEDs in the matrix are on. The sensitivity can then be set as required with the aid of the volume control, P1.



## ELECTRONICS SCENE

### Keep Confidental Meetings Private

A scanning radio receiver designed to detect even the most advanced eavesdropping devices has been developed by Audiotel International.

This effective counter-surveillance device, named 'Scanlock 2000', detects not only hidden radio microphones, but also listening devices that use mains cables for transmission. It sweeps radio frequencies from 10 MHz to 4 GHz and detects AM, FM and SC (sub-carrier 20 kHz to 130 kHz) transmissions originating within the room. It can be used to 'sweep' a room before a meeting and then left 'on guard' during the meeting to detect any 'bugs' that become active as a result of timers or remote actuation.

In automatic mode, Scanlock locks to the strongest local transmitter; even the lowest-powered 'bugs' in close proximity will have a higher signal strength than more powerful but more distant legitimate transmitters. Scanlock locks to a signal in under half a second. In manual mode, the entire RF spectrum can be covered far more rapidly than on a conventional radio receiver owing to the advanced manner in which the unit monitors harmonically related frequencies simultaneously.

When a signal is located, the program being carried may be monitored directly on headphones or a built-in speaker; confirmation that the transmission is from a 'bug' nearby is obtained by the Scanlock emitting a 2 kHz tone and demonstrating the presence of this tone in the transmission being monitored.

### Industry Prepares for the Videophone

INMOS expects to be the first manufacturer to offer a computer device that will allow videophones to meet international standards due to be agreed this year.

The company has recently launched its new high-performance signal processing device that is capable of rapid and high-quality image data compression and decompression for image handling applications such as videophones and teleconferencing systems, photographic-quality colour facsimile, interactive video disc systems and high-definition colour images in desktop publishing.

Such systems require very high levels of 'number crunching' to solve the problems of digital data storage, communication and reproduction. The new INMOS discrete cosine transform (DCT) image processor is capable of 320 million operations a second, providing computational power for the real-time DCT calculations needed to compress video signals at data rates of up to 20 MHz.

The INMOS IMS-A121 DCT image processor is housed in a 44-pin PLCC package. Its initial price for quantities of 1000 or more will be £47, but once volume production has started later this year, the price is expected to fall to less than half this figure.

### Computer-displayed Artwork

A user's own artwork can be employed as the basis of a computerized display and information system for educational, training or promotional purposes by means of a hardware and software package from Crystal Presentations.

The 'Electronic Display Unit' consists of a touch-sensitive A1 pad (594×841 mm) to which the artwork is attached and then connected to an IBM PC or compatible.

The system is used, for instance, by British Aerospace to familiarize pilots and maintenance engineers with the warning annunciators on aircraft-flightdeck overhead panels.

# CAPACITORS FOR RF APPLICATIONS

**A brief overview of currently available devices in an electronic component group whose significance for telecommunications and radio circuits is often grossly underestimated.**

Today's electronic component market offers a vast range of capacitors for use in high-frequency circuits. Many design engineers and home constructors are, therefore, often faced with a real dilemma when it comes to choosing and mounting the right capacitor in the right place. In many cases, parts lists of construction projects will provide the value of the capacitor, but references like 'PTFE foil trimmer', 'coffin-type leadless ceramic', 'ceramic NP0', 'tubular', and many more, may not be familiar.

## Ceramic capacitors

These small devices are probably the best known types for use in RF circuits because they are cheap and have been with us for many years. The lead spacing is usually 2.5 or 5 mm for the modern disc and plate types, of which some have rims below the capacitor body to facilitate their fitting at a uniform height above the PCB surface (Stettner's 'Hot Pants™' types). Values range from 0.68 pF to 100 nF. Types with values greater than about 4.7 nF usually have a working voltage of 63 V, or 12 V for sub-miniature types. Tubular capacitors are rapidly becoming obsolete and are not recommended for new designs.

The blue 'Sibatit' types from Siemens are often used as decoupling capacitors in video and digital circuits because they line up nicely with DIL ICs, require little space and offer excellent RF and pulse characteristics. Sibatit capacitors must not be confused with MKT types, which are also blue but use a multi-layer polytheraphtelate dielectric structure. Popular values of Sibatit capacitors are 10 nF, 47 nF and 100 nF. The terminal pitch is usually 5 mm, and the maximum working voltage 63 V.

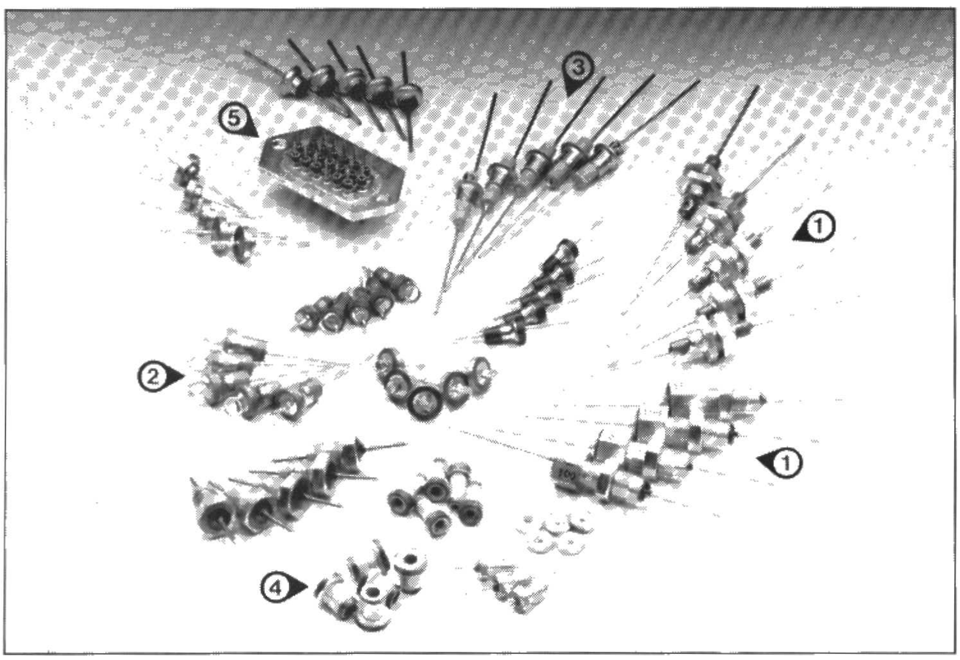
Unfortunately, the value of the common ceramic plate and capacitor is often not immediately evident from the print on the device, so that a capacitance meter is required in case of doubt. Space restrictions do not allow, say, 220p or 220pF to be printed on the capacitor body. Instead, the value is printed as, for instance, 'n22', avoiding problems with a (tiny) decimal point, or confusion with trailing zeroes. Similarly, a value of, say, 47 nF, is often printed as '473', meaning '47' with three zeroes: 47,000 pF.

The temperature coefficient of a ceramic capacitor is indicated by the coloured band at the top of the body. Although many manufacturers deviate from the standard, a zero-coefficient, or 'NP0', capacitor is generally marked with a black band. NP0 capacitors are often used in oscillators to prevent temperature changes causing frequency drift.

## Filters

Polystyrene and polypropylene capacitors are fine for filters in audio and video equipment and radio circuits (but not RF power amplifiers) for up to 30 MHz. Also known as 'styroflex' types (a trademark of Norddeutschen Seekabelwerke AG, of Nordenham, Federal Germany) manufactured by Siemens, these capacitors are usually supplied with relatively thin, axial leads, although radial, plastic encapsulated types are also available. The black or red band on the white, greyish or silver-coloured capacitor body indicates the terminal that is connected to the outer foil layers. If applicable to the circuit, this terminal must be connected to ground to provide a screening function. Special polystyrene types are available as parallel pairs for calibration purposes. These capacitors have a non-standard, but accurately defined, value at a tolerance of 0.5% or better, which makes them eminently suitable for calibration of inductance and capacitance meters.

The differences between polystyrene and polypropylene capacitors mainly entail the loss factor, permissible humidity, self-resonance frequency and insulation resistance. For most practical applications, however, polystyrene and styroflex types are all right. The version with the thin axial leads will be superseded by plastic encapsulated radial types because these have a fixed size and lead spacing and are, therefore, easier to handle in automated PCB solder machines.



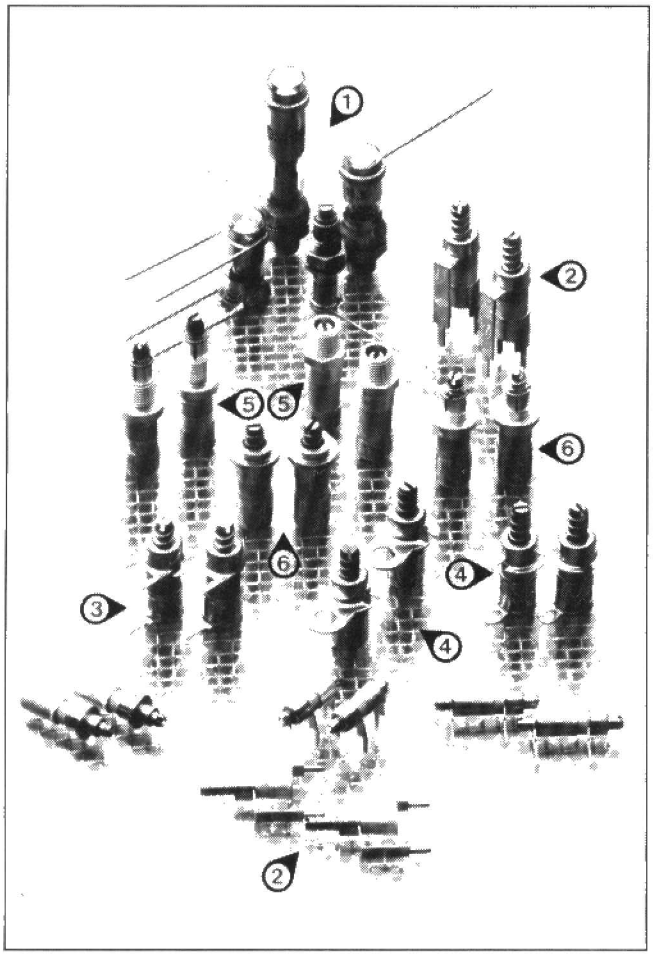

**Feedthrough capacitors: ① screw-type heavy duty; ② solder type with eye; ③ axial solder type; ④ low-capacitance feedthrough; ⑤ multiple feedthrough.**

## Trimmers

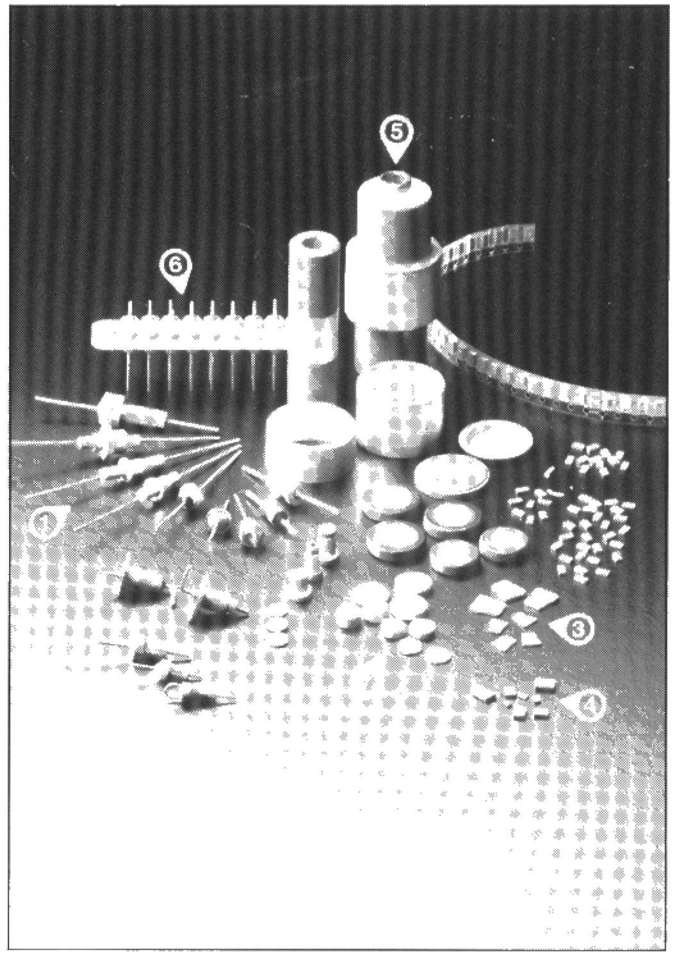
Compression and mica foil trimmers are generally not used at frequencies above 30 MHz or so because their losses increase rapidly, and the Q (quality) factor drops to an unacceptable level. Further, their minimum capacitance is not low enough for tuned circuits in the VHF, UHF and SHF range.

Ceramic trimmers suitable for frequencies up to 500 MHz are available up to a capacitance of 100 pF. The maximum capacitance is, however, nearly always less important than the minimum capacitance, which is typically 15 to 30% of the maximum value.

PTFE foil trimmers are often preferred to ceramic types because the insulating foil layers between the rotor and the stator are transparent. This allows the set capacitance to be deduced readily from the position of the grounded rotor blades with respect to the stator. The foil trimmers from Valvo (Philips Components) are colour-coded to indicate the maxi-



Tubular trimmers for use in SHF circuits. ① glass dielectric types; ② PCB mount types (vertical); ③ PCB mount types (horizontal); ④ chassis mounting types with screw and solder connection; ⑤ single-hole mounting types.



A selection of RF capacitors. ① screw type feedthrough types; ② solder type feedthrough; ③ coffin-type leadless ceramic; ④ leadless chip; ⑤ heavy-duty ceramic holders for use in high-power RF amplifiers; ⑥ feedthrough connection block.

mum capacitance: grey (5.5 pF); yellow (10 pF); green (20 pF); grey (40 pF); red (65 pF) and purple (100 pF). The difference between the 5.5 pF and 40 pF type is immediately apparent from the size.

Tubular ceramic or glass trimmers for chassis and PCB mounting are used in SHF circuits where line inductors or etched striplines are to be tuned at minimum loss. Well-known manufacturers of high-quality air, glass and ceramic tubular trimmers are Johansson, Tronser, Arco, Sky, Jackson and Stettner.

## Decoupling capacitors

Disc- and coffin shaped leadless ceramic capacitors are not so familiar among constructors with little RF experience. These capacitors are characterized by very low inductance, which is mainly by virtue of the absence of connecting wires. Instead, metallized surfaces are used for soldering at both sides of the device. The coffin-shaped capacitor (called *trapezoidal* by Stettner) is inserted and clamped in a 4×0.7 mm PCB slot that keeps the device in place during the soldering process. The printed circuit board tracks to the capacitor should be relatively wide to allow the low inductance of the device to take effect. The value of the coffin-type capacitor is printed on one metallized side. Values range from about 5 pF to 2 nF at a maximum working voltage of 63 V.

Disc-shaped leadless capacitors are more difficult to handle because they must, in general, be soldered on to a flat surface. It is difficult, however, to heat one side of the capacitor, and the surface it will rest on, simultaneously. To ensure a good solder joint and sufficient adhesive strength, a small hole is often drilled in the surface, and some hot tin is applied through it from the other side. In general, both the coffin and the disc-shaped leadless ceramic capacitor should be handled with utmost care because they are relatively brittle devices.

## The smaller the better?

The increasing use of surface-mount assembly (SMA) components has not gone past unnoticed in the RF engineering field where the size of parts has always been a crucial factor. Not surprisingly, therefore, many RF circuits are currently produced in SMA technology. For capacitors in RF circuits, this brings many benefits because SMA parts generally have a lower stray inductance than their normal-size equivalents.

SMA capacitors must not, however, be confused with chip types, which have been around for many years and offer even lower stray inductance.

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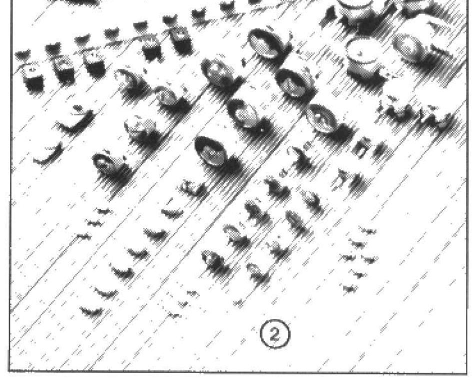
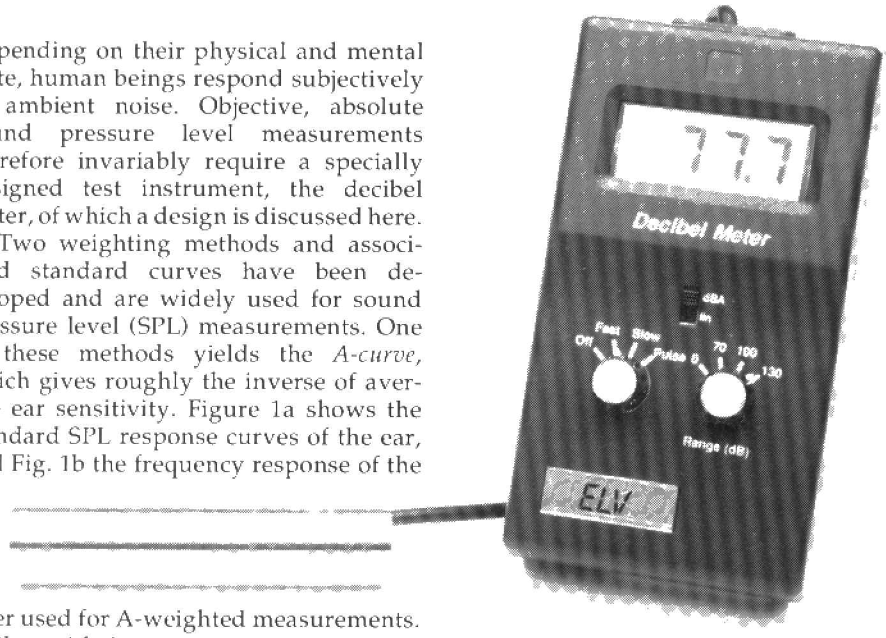


Trimmers. ① tubular trimmers; ②ceramic trimmers. The tape at the centre contains SMA trimmers.

# DECIBEL METER

**This portable instrument, designed by ELV GmbH, gives an accurate indication of the sound pressure level (SPL). The three SPL ranges, three response modes, and linear or A-weighted filtering provided by the meter enable many types of measurement to be carried out, from the tracing of ambient noise sources to establishing the sensitivity of a loudspeaker.**




Depending on their physical and mental state, human beings respond subjectively to ambient noise. Objective, absolute sound pressure level measurements therefore invariably require a specially designed test instrument, the decibel meter, of which a design is discussed here.

Two weighting methods and associated standard curves have been developed and are widely used for sound pressure level (SPL) measurements. One of these methods yields the *A-curve*, which gives roughly the inverse of average ear sensitivity. Figure 1a shows the standard SPL response curves of the ear, and Fig. 1b the frequency response of the filter used for A-weighted measurements. A filter with A-type response is incorporated in the present decibel meter to enable it to give readings which are meaningful in terms of human response to sound pressure levels and variations thereof.

In a number of cases, it is useful to perform linear, that is, unweighted, SPL measurements. The SPL sensitivity of a loudspeaker, for instance, is established in this manner across a frequency range of 20 Hz to 20 KHz, with the decibel meter set to linear (unfiltered) response.

## Controls and operation

The decibel meter has two rotary switches and one slide switch on the front panel. All functions of the instrument are selected with these three controls.

The slide switch selects linear (LIN) or A-weighted (dBA) response. In general, A-weighting is used whenever human response to sounds or noise is involved. This is so in practically all cases where SPL measurements are used for determining the degree of ambient noise. The unweighted type of measurement, on the other hand, is commonly used for strictly technical work such as the design of a loudspeaker enclosure.

The right-hand rotary switch on the front panel forms the range selector. When turned fully counter-clockwise to position 0, the analogue-to-digital (A-D) converter and the associated preamplifier in the instrument are tested. The display shows a value between 00.0 and 00.4 (with the left-hand rotary switch set to position F). Higher values on the display indicate a fault in the decibel meter. Position 70 dB allows SPL measurements between 40 and 70 dB to be taken; position 100 dB, measurements between 70 and 100 dB; and position 130 dB, measurements between 100 and 130 dB. Each range can be used for readings of 10 dB below the minimum value or above the maximum value, with only marginally lower accuracy. The

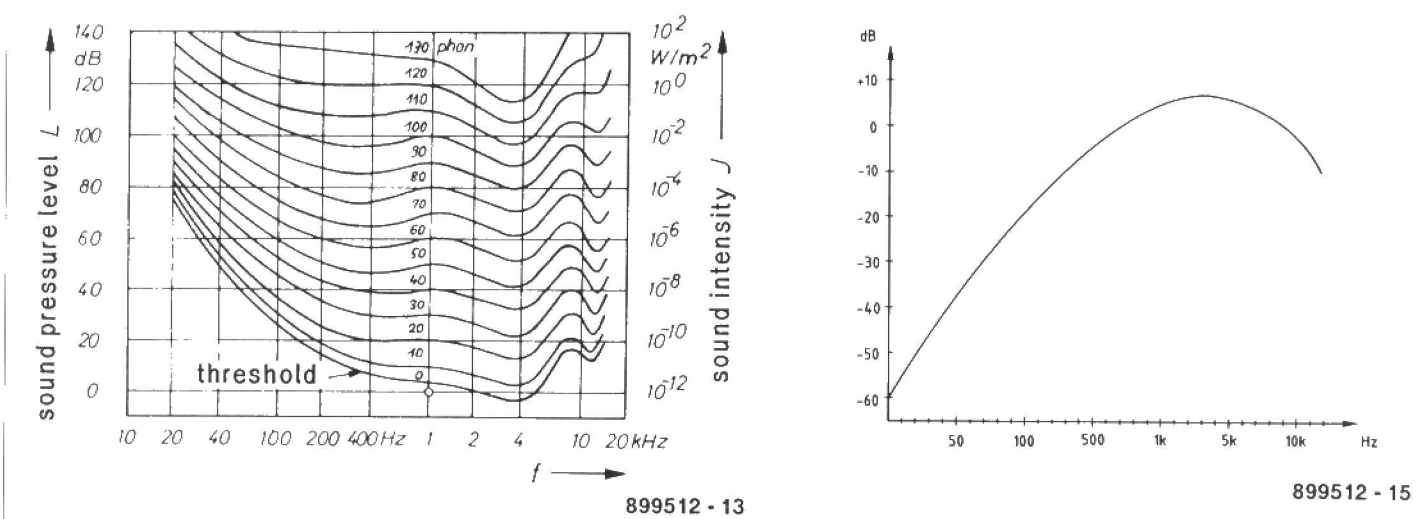



Fig. 1. **Equal loudness contours adopted for aural response testing (left) and the A-weighted filter characteristic derived from these (right).**

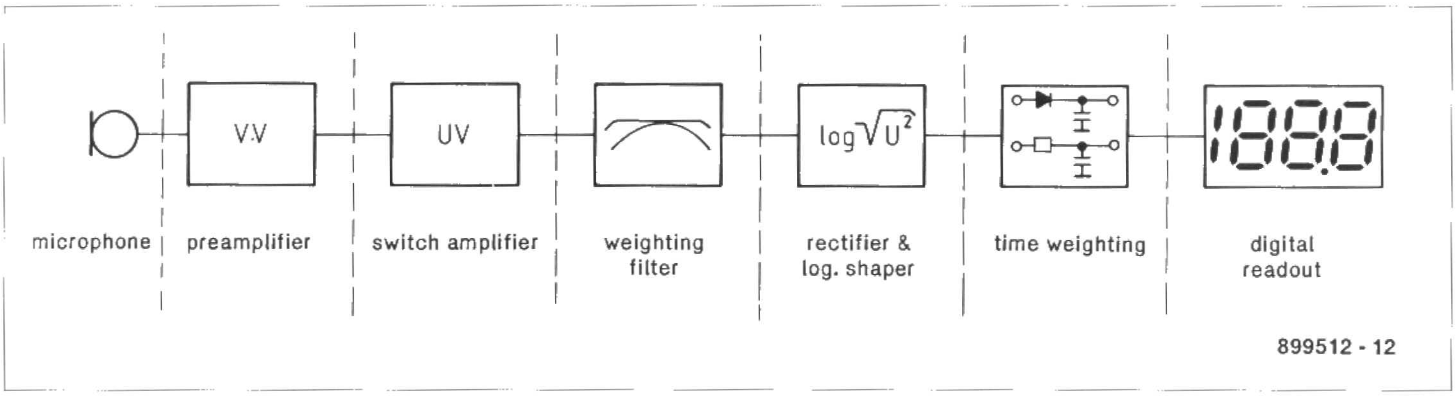



**Fig. 2. Block diagram of the decibel meter.**

measurement ranges thus have a sufficiently large overlap.

Accuracy of the decibel meter within the above three ranges (effectively from 40 to 130 dB) is 0.5% typically. In most cases, the usable low and high limit is 35 dB and 135 dB respectively, which corresponds to a dynamic range of no less than 100 dB, or a ratio 1:100,000.

The accuracy of the instrument is also determined by the user who must select a range suitable for the relevant measurement before this is actually performed. The reading obtained during this test should fall within the limits of the selected range. If necessary, a higher or lower range is selected. Apart from small deviations from the minimum and maximum levels of a particular range, readings that fall outside the scale should not be used. This is because measurement errors rise rapidly for readings more than 10 dB below the minimum or above the maximum value of the selected range.

The left-hand rotary switch selects the speed of the measurement, and also serves to switch the instrument on and off.

In many cases, noise levels fluctuate and give rise to incorrect readings if the decibel meter is not set to the appropriate speed. The present instrument offers three speed ranges for SPL measurements.

**Slow** (switch position S)
A time constant of 1 s is used to ensure a stable reading even when the SPL varies considerably. The slow meter response is particularly useful for long-term measurements to give average sound levels. Fast changes in the SPL values are averaged out, while pulses are effectively eliminated.

**Fast** (switch position F)
In this mode, a time constant of 125 ms is used to enable fast SPL changes to be recorded reliably. The display reading settles relatively quickly, and follows rapid SPL variations. This type of meter response is obligatory for certain peak-SPL measurements.

**Pulse** (switch position P)
Two time constants are used in this mode. Particularly suited to pulse and intensity measurements, this range uses a short rise time of 35 ms and a long decay time of 10 s. It should be noted that meter overflow may occur readily when the range setting of the instrument is changed. If this happens, it is necessary to wait 10 s until the measured value has dropped to a value lower than the expected SPL.
In practice, the selection of a particular speed response is determined by the application and the required read-out (peak value, average value, or fluctuations).

## Block diagram

The block diagram of Fig. 2 shows the essential configuration of the decibel meter. The sound pressure to be measured is applied to a high-grade electret microphone. This device, a Type KE4-211-2 from Sennheiser, is constructed in back-electret technology, and has an internal impedance converter. The element is housed in a TO-18 style transistor enclosure, and provides a good bandwidth (20 Hz to 20 kHz), a high signal-to-noise ratio, and a large dynamic range (approx 35 dB to 135 dB). The back-electret technology ensures low sensitivity to vibration, and eliminates mechanical resonance. All these features make the KE4-211-2 an excellent choice for application in the present decibel meter, whose overall performance is certainly degraded if an electret microphone element with mediocre specifications is used.

The transducer and the associated preamplifier are housed in a tube that forms the actual microphone. This construction is necessary to ensure that the instrument can measure voltages of the order of 10 µV in the most sensitive range.

An approximately 1 m long cable connects the microphone to the decibel meter. The use of separate units affords flexibility for the measurements, and also prevents the main instrument from introducing interference in the area to be examined. The preamplifier is followed by the sensitivity selector which offers three ranges as discussed above.

The next stage is a filter that allows a selection to be made between linear (flat) frequency response, and weighted response on the basis of the A-curve.

The amplified AF voltage is applied to a quadrature rectifier with a logarithmic shaper at the output. The rectified voltage must be converted from linear to logarithmic to enable the digital voltmeter (DVM) section to be driven in accordance with the required scale curve. The time weighting section between the output of the logarith-

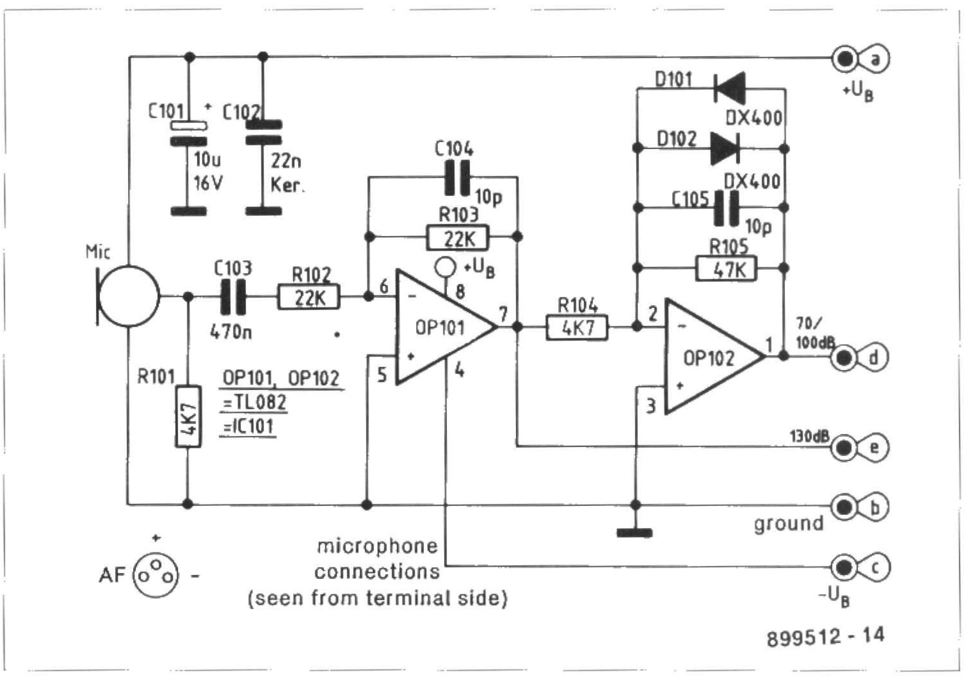



**Fig. 3. The microphone preamplifier is a two-stage design that uses a precision electret microphone from Sennheiser.**

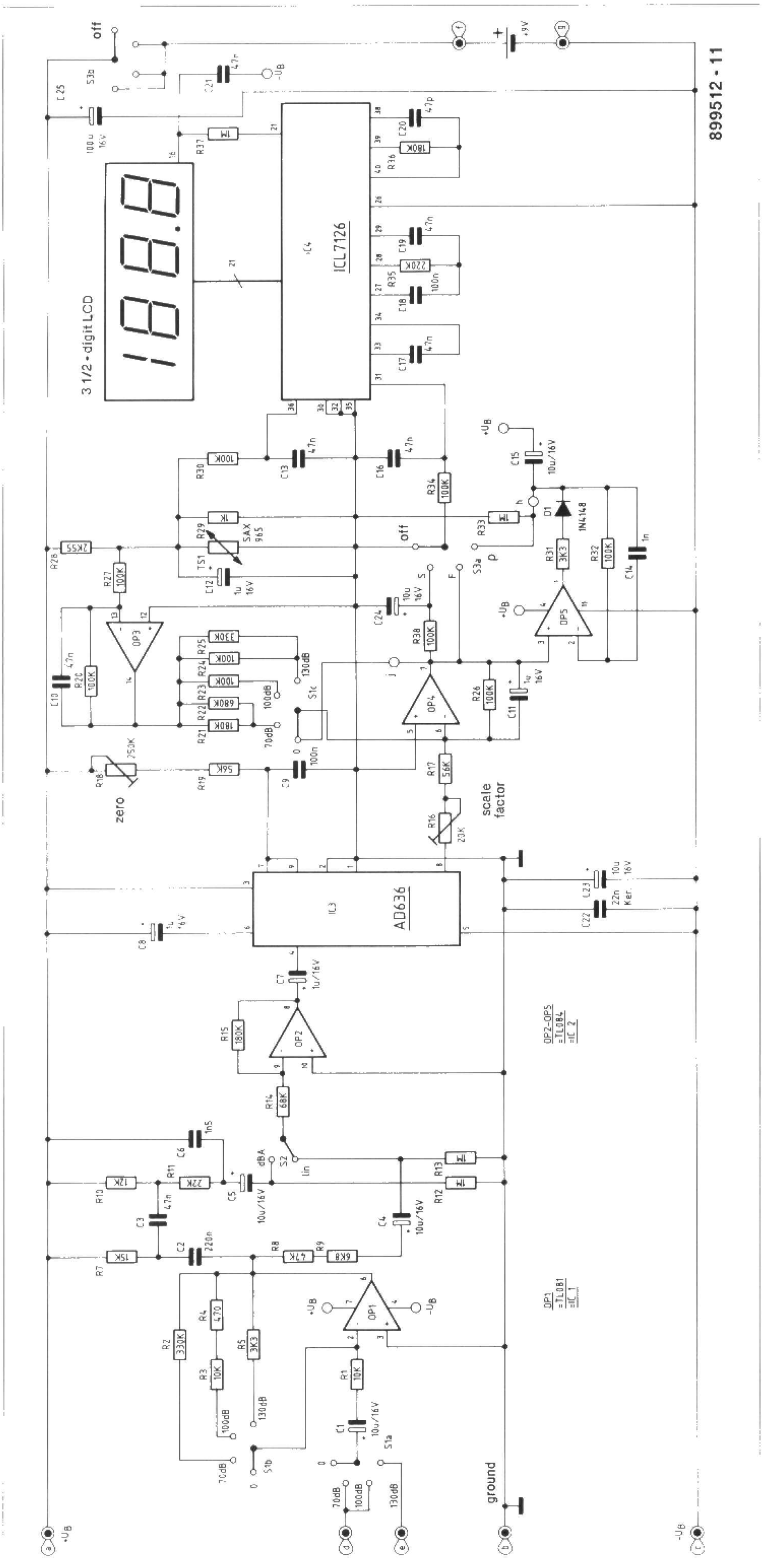



**Fig. 4. Circuit diagram of the main instrument. Note the use of temperature compensation that acts on ADC IC3 and display driver IC4.**

mic shaper and the input of the DVM allows selection of peak level measurement or a number of time constants for signal integration.

Finally, the DVM and associated display give a direct SPL reading in dB.

## About the circuit

The circuit diagram of the microphone, i.e., the pressure transducer and associated preamplifier, is given in Fig. 3. The microphone element is, in principle, a hyper-sensitive pressure transducer that records air pressure changes caused by sound sources, and converts these changes into an alternating voltage. This signal is applied, via $C_{103}$ and $R_{102}$, to an impedance converter built around opamp $OP_{101}$. The gain of this stage is set to unity with resistors $R_{103}$ and $R_{102}$, while the phase shift is 180 degrees. The output signal of $OP_{101}$, available at pin 7, is used for the 130 dB range of the instrument. The relevant connection is made via a PCB terminal marked 'e'.

The two ranges with higher sensitivity require a further amplification of 10 times, which is achieved with $OP_{102}$ and $R_{105}$-$R_{104}$. Two Schottky diodes with low reverse current, $D_{101}$ and $D_{102}$, are included in the feedback path of the second opamp to prevent this affecting the operation of the first stage when, for instance, the output of $OP_{102}$ reaches the clipping level as a result of high sound pressure levels that may be measured in the 130 dB range. In the lower ranges (70/100 dB) the signal levels are so low that the diodes have no effect.

Capacitors $C_{101}$ and $C_{102}$ stabilize the amplifier and prevent it from oscillating. Resistor $R_{101}$ forms the load required for terminating the microphone element.

The power supply for the microphone circuit is applied via terminals 'a' (positive battery voltage) and 'c' (negative battery voltage). Terminal 'b' connects ground of the microphone to the virtual ground level created in the main instrument. As already discussed, terminals 'd' and 'e' form the signal outputs. The microphone is connected to the main instrument by a 1 m long 4-way screened cable. The screening is connected to point 'b' on the preamplifier PCB.

The circuit diagram of the main instrument is given in Fig. 4. Rotary switch $S_{1a}$ carries the AF signal from the microphone unit to the input of the range amplifier built around $OP_1$. The required amplification factor is selected with $S_{1b}$. The ratio of the selected resistor in the feedback network to $R_1$ determines the amplification. Capacitor $C_1$ decouples the direct voltage.

Depending on the position of slide switch $S_2$, the AF signal supplied by $OP_1$ is fed either to the A-weighting filter, $C_2$-$C_3$-$C_6$-$R_7$-$R_{10}$-$R_{11}$, or, via $R_8$-$R_9$, to inverting amplifier $OP_2$, whose output signal is capacitively coupled to the input of $IC_3$. Capacitors C4, $C_5$, and $C_7$ serve to decouple the direct voltages.

Circuit $IC_3$ is an analogue-to-digital converter Type AD636. Originally designed as an rms-converter (Ref. 1), this chip has an on-chip linear-to-logarithmic converter with a dB output. The dB converter is driven by the true-rms converter, and supplies an output voltage that corresponds to the logarithm of the rectified value. The logarithmic voltage available at pin 8 of $IC_3$ is applied to an inverting amplifier built around $OP_4$. The amplification of this stage, which determines the scale factor, is set by $R_{16}$.

The position of time weighting selector $S_{3a}$ determines whether the output voltage of $OP_4$ is obtained direct (fast response), via 1-s time constant $R_{38}$-$C_{24}$, or via a peak rectifier built around $OP_5$. In the latter circuit, the charge time constant is determined by $R_{31}$-$C_{15}$, and the discharge time constant by $R_{33}$-$C_{15}$.

The selected voltage is fed to the input of A-D converter $IC_4$, an ICL7126. This IC converts the potential difference that exists between its pins 30 and 31 into a corresponding digital value, which is shown on an LC-display. The associated reference voltage exists between pins 35 and 36. An internal reference source supplies a voltage that is available at pin 32. This voltage is typically 2.8 V lower than the positive supply voltage (9 V in this case).

A special component in the display driver circuit is temperature sensor $Ts_1$, a Type SAX965. Resistors $R_{28}$ and $R_{29}$ cause the regulation effect of the temperature sensor to be such that the reference voltage applied to pin 36 of $IC_4$ varies in accordance with the temperature of $IC_3$, and therefore with the voltage drift at the output of this ADC. This temperature compensation arrangement is required to ensure the accuracy of the voltage at the dB output.

The compensation currents for the three measurement ranges are changed to the same extent as the reference voltage, but with a different sign. Inversion is accomplished by $OP_3$, which is connected to the reference voltage via $R_{27}$. At an ambient temperature of 25°, $Ts_1$ drops a reference voltage of about 470 mV.

The values of $R_{21}$ and $R_{25}$ are geared to achieve a level shift at the output of $OP_4$ that results in a difference of 30 dB between the ranges.

The circuit is nulled with the aid of preset $R_{18}$, which feeds an additional (offset) direct current into the lin-log converter. The null adjustment is actually carried out by selecting a measurement value towards the end of the range, and adjusting $R_{18}$ to obtain a reading that corresponds to the applied input signal. The setting up of the decibel meter will be reverted to.

## Construction

The complete circuit is built on two relatively small, single-sided, printed-circuit boards. Populating these in accordance with the Parts List and the overlays in Fig. 7 is not likely to present difficulties.

The low-profile components are mounted first, followed by the higher components. Three parts are fitted horizontally: $C_{101}$ on the preamplifier board, and $C_{12}$ and $Ts_1$ on the main board.

Although it is, of course, best to mount the temperature sensor such that it has direct thermal contact with $IC_3$, acceptable compensation is also achieved by the component placement on the PCB. This is so because the circuit is enclosed in a cabinet and will normally work at a fairly even ambient temperature. If the temperature drops suddenly by 10 degrees or more, about half an hour should be allowed for the instrument to settle at the new ambient temperature.

The following components are fitted at the track side of the main PCB: $R_{35}$, $C_1$, $C_{13}$, $C_{16}$–$C_{20}$ and $C_{25}$. The terminals of $C_{25}$ are soldered direct to pins 4 (+ wire) and 11 (– wire) of $IC_2$.

Fit soldering pins in the six holes for the slide switch, $S_2$, and cut the pins to about 2 mm. The slide switch is secured by soldering its terminals to the pins.

The correct mounting height of the LC display is ensured with the aid of a 40-pin IC socket, which is first cut lengthwise. Next, the two 20-pin rows are soldered on to the PCB. The drawing in Fig. 6 shows how a second IC socket is stacked on to the one on the PCB. This second socket is also cut in two so that it can receive the pins of the LC display.

The solder terminals marked'h' on the main PCB are interconnected by a 35 cm long flexible insulated wire installed at the component side. Similarly, points 'j' are interconnected by a 25 cm long wire.

The wires of the battery clip are soldered to terminals 'f' (+; red) and 'g' (–; black).

The microphone wires, which are left at their original length, are inserted from the component side in the holes provided in the preamplifier board. The wires are then soldered at the track side, and afterwards bent at right angles where they leave the printed-circuit board at the component side. This ensures that the microphone points forward so that its face can protrude from the tube to be fitted later.

The preamplifier board is ready for mounting inside the approximately 100 mm long metal tube after the 4 wires and the screening of the cable to the main instrument have been connected to the appropriate solder terminals 'a' through 'e'. The small PCB is then inserted into the tube to a point where the top of the microphone element protrudes about 3 mm from the front side of the tube. A heavy-duty soldering iron is then used to solder a part of the ground surface of the preamplifier PCB to the inside of the metal tube. This solder joint should be made quickly and as far as possible to the rear of the tube to prevent overheating the sensitive element at the front side.

The remainder of the construction entails connecting the 4 wires and the cable screening to the main instrument, and installing this in the enclosure provided with the kit.

## Setting up

An ammeter is inserted between the completed instrument and the 9 V battery. The left-hand rotary switch is set to position OFF, the right-hand switch to position 0, and the slide switch to position LIN. The current must be nought, and the display blank.

The left-hand rotary switch is turned to position F (fast). The resulting current is between 4 and 10 mA (6 mA typ.). The display should indicate a value between 00.0 and 00.4. If not, the circuit should be examined for faulty solder joints and incorrectly placed components, particularly in the area around $IC_2$ and $IC_4$.

The right-hand rotary switch is advanced to the 70 dB position, when the operation of the instrument can be verified by saying a few words into the microphone.

Now connect a sine-wave generator capable of supplying a 1 kHz output signal at low distortion to the temporarily used adjustment circuit shown in Fig. 5. The wiper of the 1 kΩ preset is connected to junction $R_{101}$-$C_{103}$ on the preamplifier board by a short length of screened wire (the screening is connected to the ground plane on the preamplifier PCB). The microphone terminal to the above junction is temporarily removed. The sine-wave generator must have a separate power supply.

| Step | Range (dB) | 1 kHz test signal ($U_{rms}$) | $R_x$ (Fig. 5) | Indication | Adjust with |
|---|---|---|---|---|---|
| 1 | 100–130 | 631 mV ≡ 63.1 Pa | wire | 130 dB | $R_{18}$ |
| 2 | 100–130 | 20 mV ≡ 2 Pa | 39k | 100 dB | $R_{16}$ |
| 3 | 100–130 | 631 mV ≡ 63.1 Pa | wire | 130 dB | $R_{18}$ |
| 4 | 100–130 | 20 mV ≡ 2 Pa | 39k | 100 dB | $R_{16}$ |
| 5 | 70–100 | 20 mV ≡ 2 Pa | 39k | 100 dB | — |
| 6 | 40–70 | 631 μV ≡ 0.063 Pa | 1M | 70 dB | — |

**Table 1. The setting up procedure**

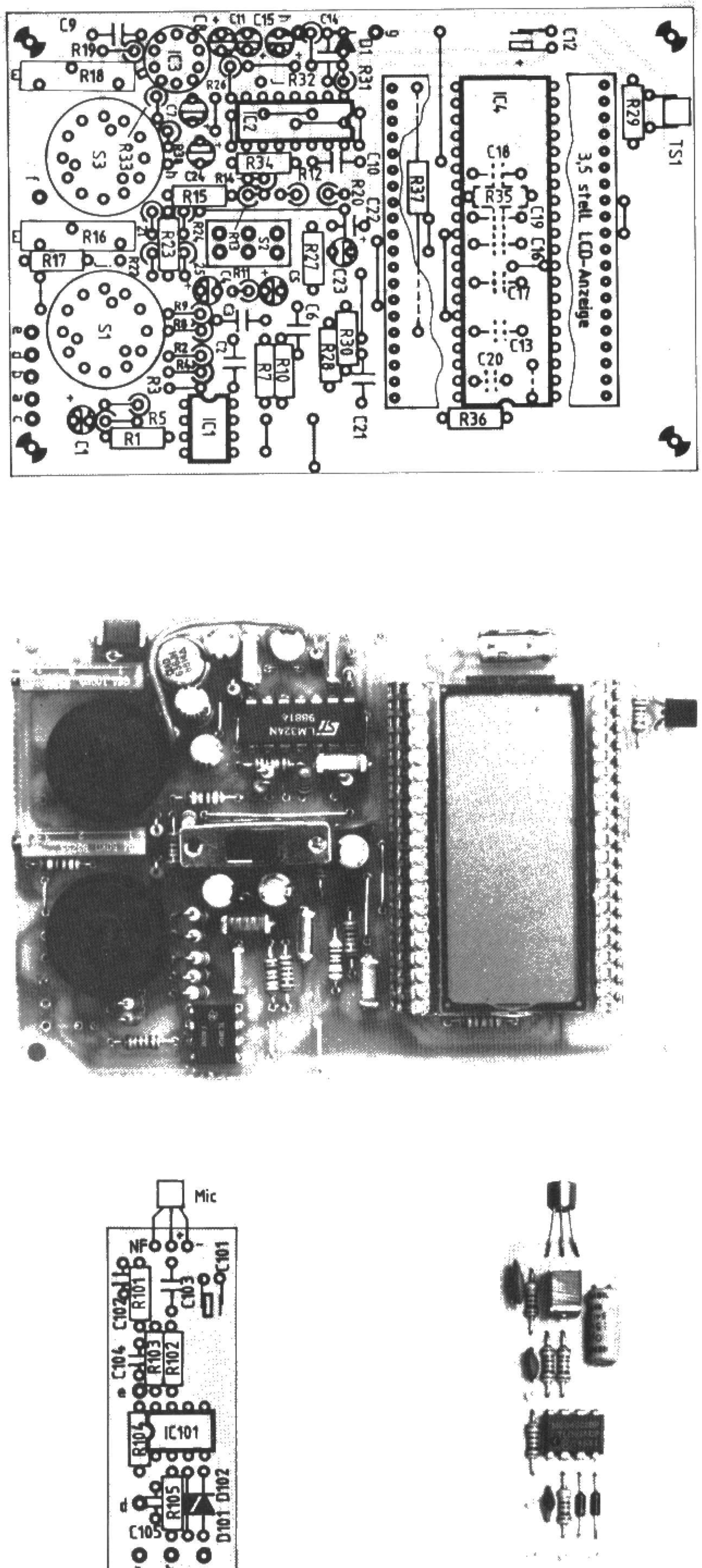



Fig. 5. Component overlays and top views of the completed printed-circuit boards.

**Parts list**

MAIN INSTRUMENT

**Resistors:**

$R_4$ = 470Ω
$R_{29}$ = 1k
$R_{28}$ = 2k55
$R_5$;$R_{31}$ = 3k3
$R_9$ = 6k8
$R_1$;$R_3$ = 10k
$R_{10}$ = 12k
$R_7$ = 15k
$R_{11}$ = 22k
$R_{17}$ = 27k
$R_8$ = 47k
$R_{14}$;$R_{19}$ = 68k
$R_{20}$;$R_{23}$;$R_{24}$;$R_{26}$;$R_{27}$;$R_{30}$;$R_{32}$;$R_{34}$;$R_{38}$ = 100k
$R_{15}$;$R_{21}$;$R_{36}$ = 180k
$R_{35}$ = 220k
$R_2$;$R_{25}$ = 330k
$R_{22}$ = 680k
$R_{12}$;$R_{13}$;$R_{33}$;$R_{37}$ = 1MΩ
$R_{16}$ = 50k multiturn preset
$R_{18}$ = 100k multiturn preset

**Capacitors:**

$C_{20}$ = 47p
$C_{14}$ = 1n0
$C_6$ = 1n5
$C_{22}$ = 22n ceramic
$C_3$;$C_{10}$;$C_{13}$;$C_{16}$;$C_{17}$;$C_{19}$;$C_{21}$ = 47n
$C_9$;$C_{18}$ = 100n
$C_2$ = 220n
$C_7$;$C_8$;$C_{11}$;$C_{12}$ = 1µ0; 16 V
$C_1$;$C_4$;$C_5$;$C_{15}$;$C_{23}$;$C_{24}$ = 10µ; 16 V
$C_{25}$ = 100µ; 16 V

**Semiconductors:**

$IC_1$ = TL081
$IC_2$ = TL084
$IC_3$ = AD636
$IC_4$ = ICL7126
$D_1$ = 1N4148

**Miscellaneous:**

$TS_1$ = SAX965.
3½-digit LC display.
$S_1$;$S_3$= 4-way, 3-pole rotary switch for PCB mounting.
$S_2$ = miniature DPDT slide switch.
Clip for PP3 battery.
15 solder pins.
1 m 4-way screened cable.
2 40-way DIL sockets (for LCD mounting).
2 5-mm PCB spacers.
2 self-tapping screws.

MICROPHONE PREAMPLIFIER

**Resistors:**

$R_{101}$;$R_{104}$ = 4k7
$R_{102}$;$R_{103}$ = 22k
$R_{105}$ = 47k

**Capacitors:**

$C_{104}$;$C_{105}$ = 10p
$C_{102}$ = 22n ceramic
$C_{103}$ = 470n
$C_{101}$ = 10µ; 16 V

**Semiconductors:**

$IC_{101}$ = TL082
$D_{101}$;$D_{102}$ = DX400

**Miscellaneous:**

5 solder pins.
Microphone Type KE 4-211-2 (Sennheiser).
Metal tube.

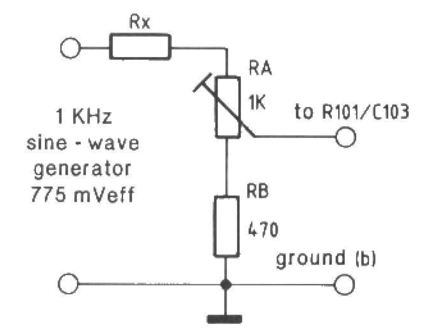



Fig. 6. Auxiliary circuit for setting up the decibel meter.

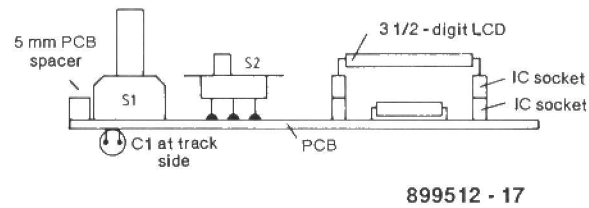



Fig. 7. Side view of the main board. Note that the LCD is mounted on to two stacked IC sockets, which have been cut lengthwise to give the correct width.

The three dB ranges of the instrument are calibrated by simulating the voltages supplied by the microphone element. The adjustment procedure is given in Table 1. The left-hand rotary switch remains in position F, and the slide switch in position LIN.

The first step in the alignment procedure is the fitting of a shorting wire in position $R_x$, and setting preset $R_A$ to an output voltage of 631 $mV_{rms}$. Preset $R_{18}$ on the main board is adjusted to achieve a display reading of 130 dB. A real zero-calibration is not possible with $R_{18}$ because the meter ranges do not reach down to 0 in actual use. The alternative is to shift the zero-point upwards and thus achieve the reading that corresponds to a particular sound level, which may be simulated by a test voltage of the correct amplitude.

The second step is the adjustment of the scale factor, with the aid of $R_{16}$, to achieve a reading of 100 dB when the preamplifier is driven by 20 $mV_{rms}$ (fit a 39 kΩ resistor in position $R_x$). This adjustment may shift the previously set null-point a little, so that the third step in the adjustment entails repeating step 1 (apply 631 $mV_{pp}$ and adjust $R_{18}$ for an indication of 130 dB).

For the fourth step in the adjustment, drive the preamplifier with 20 $mV_{rms}$ and adjust $R_{16}$ for a reading of 100 dB. The above steps are repeated until the 100 dB and 130 dB readings are obtained when the corresponding voltage is applied.

The range setting is turned to the 70–100 dB range, which is tested with the aid of a 20 $mV_{rms}$ signal. The circuit is dimensioned such that the display automatically indicates 100 dB at a maximum tolerance of 0.5 dB. No adjustments are required for this range.

The same goes for the 40–70 dB range, which is tested by applying a voltage of 631 µV. The display should read 70 dB ±0.5 dB.

Deviations exceeding the above tolerances require the resistor values to be checked carefully. If desired, the differences between the ranges can be eliminated by making small changes to $R_{23}$ (100 dB range) and $R_{21}$ or $R_{22}$ in the 70 dB range. These resistors are temporarily replaced with multiturn presets, which are adjusted for 100 dB and 70 dB when the 1 kHz test signal at the corresponding level is applied. The presets are then removed, their resistance is measured, and each is replaced by a resistor, or a combination of resistors, of the required value. In general, however, these corrections will not be necessary.

It will be clear that the previously detailed setting-up procedure yields valid results only when the microphone element has a sensitivity of 10 mV/Pa. In practice, however, a tolerance of ±2.5 dB should be taken into account. To prevent this degrading the accuracy of the decibel meter from the stated 0.5 dB, the sensitivity specifications of each microphone element have been tested by ELV. The device found in the kit is supplied with a note stating the correction factor, which can take a value between 0.7 and 1.4. The test levels given in Table 1 are to be multiplied with this correction factor.

A brief example is given to illustrate the level computations. Assuming that the microphone element has a correction factor of 1.2, the test voltage level for the first step in the alignment procedure must be

631 mV × 1.2 = 757.2 mV.

Similarly, the second step is to be performed with a test voltage of

20 mV × 1.2 = 24 mV, etc.

Finally, it should be noted that the test voltages should be measured with a voltmeter that is known to give accurate readings at a frequency of 1 kHz.

## Final assembly

The sine-wave generator and the level adjustment circuit of Fig. 5 may be disconnected if all tests and adjustments are completed satisfactorily. The screened cable is disconnected from junction $R_{101}$-$C_{103}$, and the connection to the microphone terminal is restored. Say a few words aloud to check that the instrument works in all ranges, including the dBA function and the three response modes.

Fit the PCB of the main instrument into the upper half of the hand-held enclosure, and secure it on to the battery compartment panel with the aid of two 2×10 mm self-tapping screws. Ensure the correct distance between the PCB and the inside of the enclosure by two 5 mm long spacers. At the other side, secure the PCB with a few drops of two-component, epoxy resin or super-glue.

Pass the cable to the microphone unit through a 4 mm hole in the top wall of the enclosure. After mounting the strain-relief clamp, connect the 4 wires and the screening to the appropriate points on the PCB. Care should be taken not to cause a short-circuit via the screening connection. The enclosure may now be closed with the rear panel.

The preamplifier PCB in the metal tube must be protected from ambient influences. This may be achieved by filling the tube with a suitable resin or potting compound. Great care should be taken not to spill resin over the microphone element, whose face should be temporarily covered with adhesive tape. Then carefully insert removable potting between the microphone and the inside of the tube. The compound should go a few millimeters into the tube.

Hold the tube vertically, and pour the resin in at the cable side. After about 24 hours, remove the compound around the microphone element, and carefully pour resin into the microphone side of the tube until it is full. The face of the microphone must be 2 to 3 mm above the top side of the tube. The protective tape may be removed after the resin has hardened. The decibel meter is ready for use.

**Reference:**

1. "True-RMS Meter". *Elektor Electronics* December 1986.

# PC AS TONE GENERATOR

J. Schäfer DL7PE

**A GW-BASIC program and a few modifications to the loudspeaker circuit enable any PC, whether an XT, AT or compatible, to function as a precision tone generator with a frequency range of 20 Hz to 120 kHz, with a basic sweep function as a useful option. The nice thing about this generator is that it costs next to nothing, while doubling as a frequency meter.**

The present PC tone generator, which is really a BASIC program only, is ideal for aligning a wide range of AF circuits. The frequency of the generated tone can be set accurately, so that the low-cost tone generator is suitable for applications that include the tuning of musical instruments (electronic tuning fork), the aligning of RTTY, fax and SSTV filters, and the dimensioning and testing of many other types of tone decoder. In many cases, the BASIC program obviates the use of a function generator and a frequency meter. This is of particular interest for applications in the audio range, where frequency meters are, in general, not very accurate. The PC tone generator allows AF frequencies to be defined with an accuracy of a fraction of a hertz.

## PC as tone generator

Apart from the possibilities offered by BASIC commands BEEP and SOUND, there exists a more powerful way of generating tones with the aid of a personal computer: direct control of the relevant hardware. This enables frequencies to be generated at quartz-crystal stability in the range from 20 Hz to several hundred kHz, independently of the PC's clock frequency. The lower frequency limit is fixed, but the upper limit can be made as high as allowed by the PC's internal AF amplifier. The relevant BASIC commands may be found in lines 260 through 330 of the listing.

The frequency resolution is excellent, especially in the audible range: at a basic frequency of 10 kHz, the step size is as small as 85 Hz, or 0.85%; between 2 and 3 kHz, the step size is 5 Hz (0.16%); and below 1 kHz it is 0.1 Hz (0.01%).

The PC generates the required tone via the built-in loudspeaker. This is adequate for nearly all calibration and adjustment work in the acoustic range. An oscillator, for instance, is simple to calibrate accurately by means of a beat-frequency measurement in which the PC functions as the reference. Just compare the two tones by listening to them simultaneously, and step the PC tone frequency until the difference frequency decreases. When it becomes inaudible, the frequency generated by the equipment under test can be read from the PC screen. Similarly, the PC tone generator can be set to a particular frequency, so that the oscillator can be adjusted until zero-beat is achieved.

If the generated tone is required electrically also, the loudspeaker signal must be made available on a jack socket — see Fig. 1. When a plug is inserted, the loudspeaker in the PC is automatically dis-

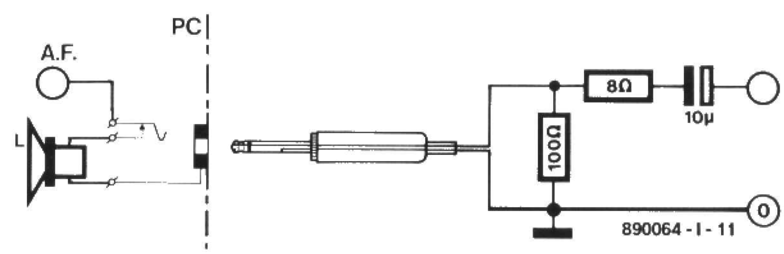



Fig. 1. Coupling out the PC's AF signal via a jack socket with a break contact.

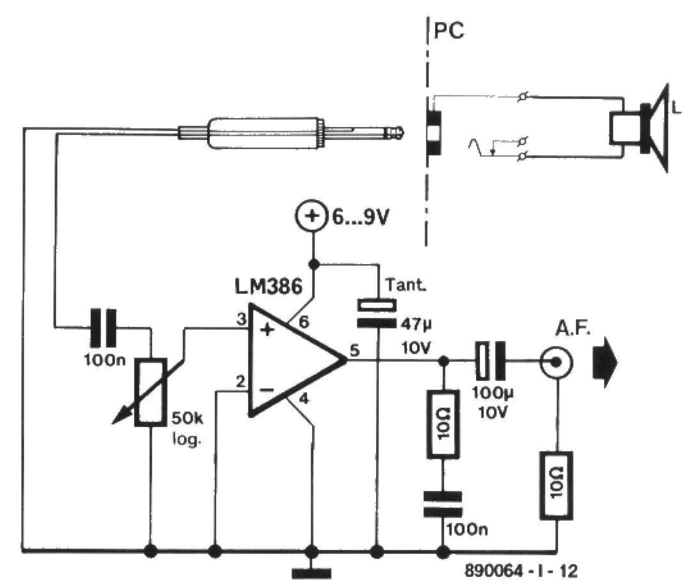



Fig. 2. The 'active' alternative: raising the PC's AF signal with the aid of a small amplifier.

abled, and the generated tone is available at a relatively low impedance. The 8 Ω resistor protects the internal AF amplifier against short-circuits. The value of this resistor must be increased as required if the internal loudspeaker is a high-impedance type. It is, of course, also possible to wire the socket such that the loudspeaker is not disabled, but taken up in series with the output signal. This arrangement obviates the use of the protection resistor. By connecting a 16 Ω resistor instead of the indicated 100 Ω type to ground, the generated tone is simultaneously audible via the internal loudspeaker.

A further possibility is shown in Fig. 2. A small amplifier with adjustable volume is connected to the jack socket. This solution is particularly useful for PCs that have an internal loudspeaker with a relatively high impedance, or one that produces insufficient output volume.

## The program

The frequency range of 20 Hz to 120 kHz is fixed in lines 180 and 200 of the the BASIC program. Keys are used to control the program:

**Key 'e':** enable tone
**Key 'd':** disable tone
**Key '+':** increase frequency by previously entered step size
**Key '–':** decrease frequency by previously entered step size
**Key 's':** terminate program

A frequency sweep is obtained by holding the + or – key — the tone frequency then increases or decreases at the previously entered step size.

## Applications

Here are a few of the many possible applications of the computer-controlled tone generator:

- test signal for aligning RTTY circuits, e.g., 1275/2125 Hz for VHF stations, and 1275/1445 Hz for SW stations.
- test signal for tone decoders
- 1,000 Hz frequency reference
- tuning fork
- elementary acoustics

Table 1 is useful for the tuning fork application because it shows the tone frequencies for three octaves.

```
10 REM: PC Tone Generator
20 REM: by DL7PE
30 CLS :KEY OFF
40 GOSUB 350
60 LOCATE 5,1
70 PRINT :PRINT :INPUT "         please enter start frequency :";FREQ
80 PRINT :PRINT :INPUT "         please enter step size :";STP
90 CLS
100 GOSUB 350
120 GOSUB 380
125 COLOR 15,0,0
128 LOCATE 9,23 :PRINT "  ┌──────────────────────────┐"
130 LOCATE 10,25:PRINT " | FREQUENCY = "; USING
"######.##";FREQ;:LOCATE 10,53:PRINT "|"
135 LOCATE 11,23 :PRINT "  └──────────────────────────┘"
140 LOCATE 10,50 :PRINT "Hz"
150 COLOR 7,0,0  :GOSUB 260
170 X$=INPUT $(1) : IF X$= "+" THEN FREQ= FREQ + STP
180 IF FREQ=>120000! THEN FREQ=120000! : GOSUB 260
190 IF X$= "-" THEN FREQ= FREQ - STP
200 IF FREQ=< 20 THEN FREQ=20 :GOSUB 260
210 IF X$= "e" OR X$="E" THEN GOSUB 640
220 IF X$= "d" OR X$="D" THEN GOSUB 660
230 IF X$= "s" OR X$="S" THEN GOSUB 660 : GOTO 250
240 GOTO 125
250 CLS : PRINT"        E N D      " : END
260 REM : generate tone
270 OUT 67,182 : REM "load 182 from timer block "
280 LET COUNT=1193180!/FREQ:REM "compute data for timer block"
290 LET CNTHI=INT (COUNT/256) :REM "most-significant byte"
300 LET CNTLO=INT(COUNT - CNTHI*256) :REM "least-significant byte"
310 OUT 67,182
320 OUT 66,CNTLO :REM "supply least-significant byte"
330 OUT 66,CNTHI :REM "supply most-significant byte"
340 RETURN
350 PRINT SPC(27);"T O N E  G E N E R A T O R  "
355 PRINT  SPC(33);"by DL7PE"
360 PRINT"________________________________________________________________
_______________________"
370 RETURN
380 LOCATE 22,1
390
PRINT"________________________________________________________________
__________________":PRINT:PRINT "key functions :     + ";
400 COLOR 0,7
410 PRINT" UP ";
420 COLOR 7,0
430 PRINT "   - ";
440 COLOR 0,7
450 PRINT" DOWN ";
460 COLOR 7,0
470 PRINT "   e ";
480 COLOR 0,7
490 PRINT" ON ";
500 COLOR 7,0
510 PRINT "   d ";
520 COLOR 0,7
530 PRINT" OFF ";
540 COLOR 7,0
550 PRINT "   s ";
560 COLOR 0,7
570 PRINT" STOP ";
580 COLOR 7,0
610 LOCATE 5,26
620 PRINT "range: 20 Hz to > 100 kHz"
630 RETURN
640 OUT 97,INP(97) OR 3 :REM turn on PC loudspeaker
650 RETURN
660 OUT 97,INP(97) AND 252 :REM turn off PC loudspeaker
670 RETURN
```

| Note | 4th octave | 5th octave | 6th octave |
|---|---|---|---|
| C | 261.6 | 523.3 | 1046.5 |
| C# | 277.2 | 554.4 | 1108.7 |
| D | 293.7 | 587.3 | 1174.7 |
| D# | 311.1 | 622.3 | 1244.5 |
| E | 329.6 | 659.3 | 1318.5 |
| F | 349.2 | 698.5 | 1396.9 |
| F# | 370.0 | 740.0 | 1480.0 |
| G | 392.0 | 784.0 | 1568.0 |
| G# | 415.3 | 830.6 | 1661.2 |
| A | 440.0 | 880.0 | 1760.0 |
| A# | 466.2 | 932.3 | 1864.7 |
| H | 493.9 | 987.8 | 1975.5 |

**Table 1. Commonly used frequencies for tuning musical instruments.**

# SIMPLE TRANSMISSION-LINE EXPERIMENTS

by Roy C. Whitehead, C.Eng., MIEE

**This article describes some simple transmission-line experiments that were developed for the UNCLE scheme. Under this scheme, which was initiated by the IEE, but later joined by other learned societies, members (usually retired) volunteer to go to schools to help teachers to bridge the gaps that exist between the academic world and the world of practical engineering.**

The material used in the experiments consisted of:

- a known length of coaxial cable of which both ends were accessible;
- a twin-beam oscilloscope;
- an HF generator with 75 Ω output;
- three 100 Ω non-inductive potentiometers;
- an ohmmeter.

## Measurement of velocity ratio

The equipment should be connected as shown in Fig. 1. Set P1 to its maximum resistance value and the two Y sensitivity controls of the CRO to produce equal values of Y sensitivity.

Set the signal generator to its minimum available frequency and note the small lateral displacement of the two waveforms. Then, increase the generator frequency, which causes the lateral displacement of the waveforms to increase, until, for the **first** time, the two waveforms are seen to be in phase. The propagation time of the cable now equals one period $t = 1/f$ of the generator output. The velocity ratio of the cable then equals

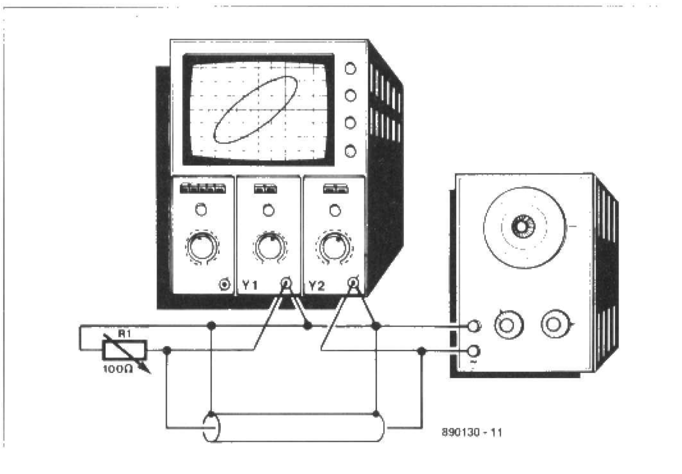



Fig. 1.

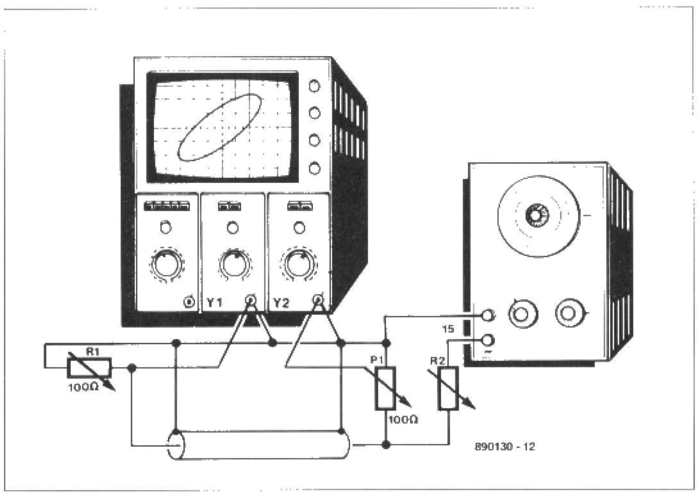



Fig. 2.

velocity in cable / velocity in free space =

= cable length in metres $\times f / (3 \times 10^8)$.

A typical value is 0.8.

Change the generator frequency to one quarter of the value used previously, which makes the line one quarter wavelength long. Adjust P1 to obtain vertical Y deflections on the CRO of equal magnitude. Disconnect P1 and measure its effective value, which is equal to the characteristic impedance, $Z_0$, of the cable.

## Measurement of attenuation/ frequency characteristics

Connect the equipment as shown in Fig. 2. Set R1 to equal $Z_0$. Potentiometer P1 has been provided with a decibel scale (which can be done with the aid of the ohmmeter). Adjust R2 to provide across the input end of the line an impedance equal to $Z_0$. If the output impedance of the generator is 75 Ω, this will be 43 Ω.

Over a range of frequencies, say from 100 kHz to the maximum at which the available equipment will operate satisfactorily, adjust P1 to produce Y deflections of equal magnitude. The cable attenuations are then equal to the attenuations of the potentiometer. The attenuation/frequency characteristic of the cable roughly follows the emperical equation

$$\text{loss} = (a\sqrt{f} + bf) \qquad [\text{dB}]$$

where b<<a so that the second term becomes significant only at frequencies above about 16 MHz, owing to the skin effect.

If a loss/frequency equalizer be designed and constructed, this may be tested in a similar manner, after which the line plus the equalizer may be tested.

Other tests may also be carried out. For instance, short-circuit the output end of the cable at Y1 and note the effects on the Y2 waveform for odd and even numbers of quarter wavelengths. Repeat this test with an open circuit at Y1.

The relationships between cable length and the frequencies at which the CRO and generator can operate satisfactorily should be noted. The shorter the cable, the higher must be the operating frequencies of the generator and the CRO.

The connectors used should preferably be coaxial, otherwise they should be short, especially when operation is at frequencies above 10 MHz.

# A NEW GENERATION OF ANALOGUE SWITCHES

by Jack Armijos and Tania Chur*

Most applications for analogue switches fall into two categories: signal routeing and signal conditioning. Different processing technologies produce switches with different characteristics. One advantage of the new CMOS analogue switches from Siliconix is that they allow you to control signals that fall anywhere between the two power supply rails. Furthermore, these

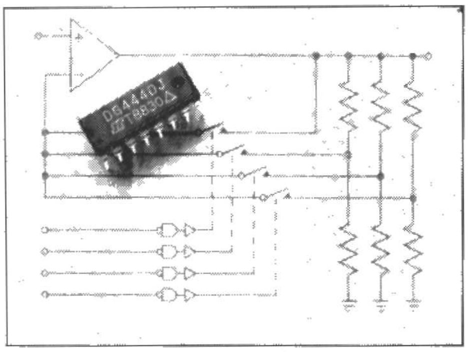


Fig. 1. The DG400 series of analogue switches.

high-performance silicon-gate ICs, the DG400 family, are pin-for-pin replacements for the popular DG200 series. They offer significantly lower on-resistance ($r_{DS(on)}$=85 Ω), lower power dissipation (35 μW), faster switching speed ($t_{on}$= 250 ns) and lower leakage current ($I_{S(off)}$ <500 pA) than the older industry-standard parts. The new devices are shown here in typical circuits, illustrating the benefits they offer.

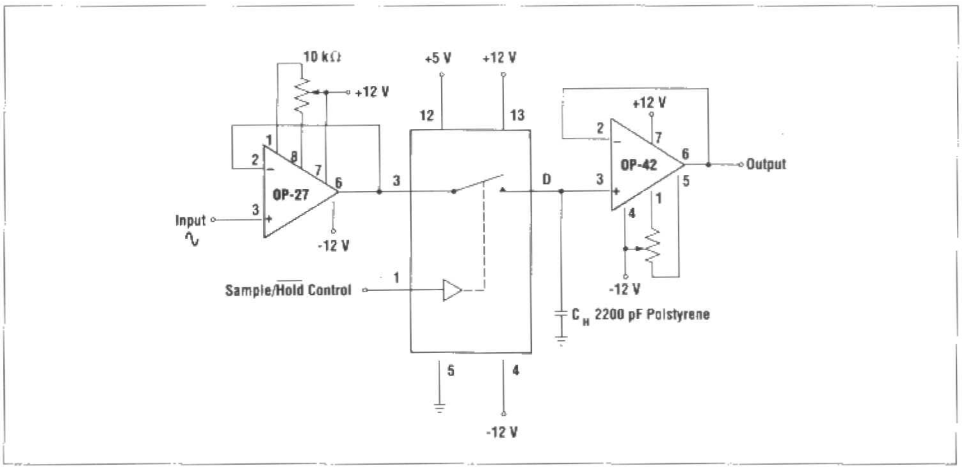



Fig. 2. Open-loop sample-and-hold circuit uses a fast analogue switch.

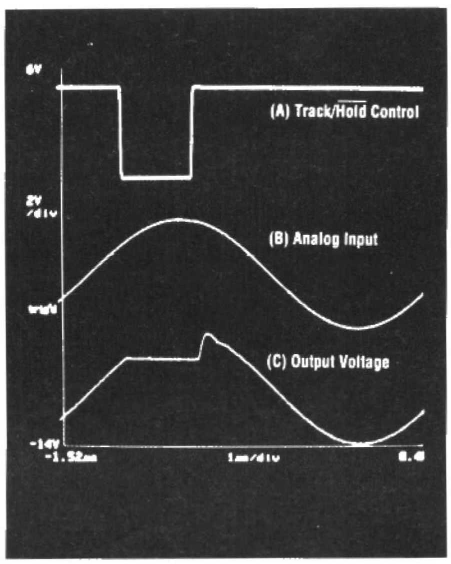



Fig. 3. A logic LOW opens the DG412, placing the circuit in the hold mode.

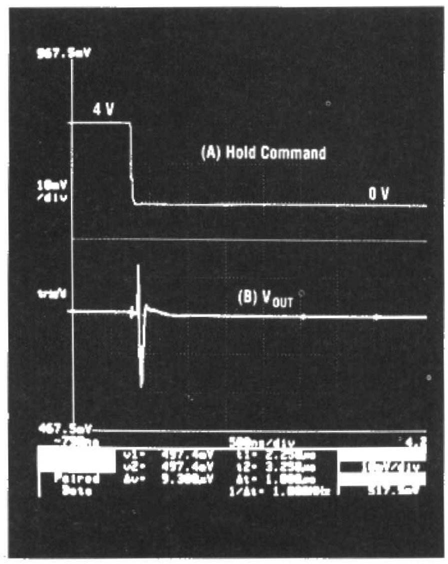



Fig. 4. $V_{out}$ showing the effects of charge injection.

## Sample-and-hold functions

In most data acquisition systems, many channels are sampled sequentially and then digitized by an analogue-to-digital converter. In choosing or designing a sample-and-hold system, speed and accuracy are the two most important considerations.

## Open-loop, cascaded-follower sample-and-hold circuit

The basic sample-and-hold circuit of Fig. 2 has unity-gain buffers to charge the capacitor without loading the signal source and to drive the next stage without changing the voltage stored. The basic operation of this circuit is illustrated in the photograph of Fig. 3. This configuration provides fast acquisition times and is good for high-speed acquisition systems.

The input buffer is chosen for low offset voltage, good slew rates, and the ability to drive the capacitive load. A polystyrene capacitor is used because of its very low dielectric absorption and low leakage. The output buffer needs to have a short settling time and very low input bias to prevent the discharge of the hold capacitor during the hold mode.

The most important switch parameters are: speed, to minimize the acquisition time (fast throughput); low charge injection, to reduce the hold step error; and low leakage, to maintain a low droop rate. The DG412 offers improvements for all three areas of performance.

The circuit shown in Fig. 2 achieved an acquisition time of under 900 ns and a droop rate of 10 μV/μs. Pedestal error was a function of analogue signal voltage. The worst-case error was 23 mV when $V_{in}$ = 5 V.

The photograph in Fig. 4 shows $V_{out}$ immediately after the hold command. In this case, $V_{in}$ = 0.5 V. Note the upset caused by the charge injection of the switch when it opens, the offset error that

* The authors are with Siliconix Ltd.

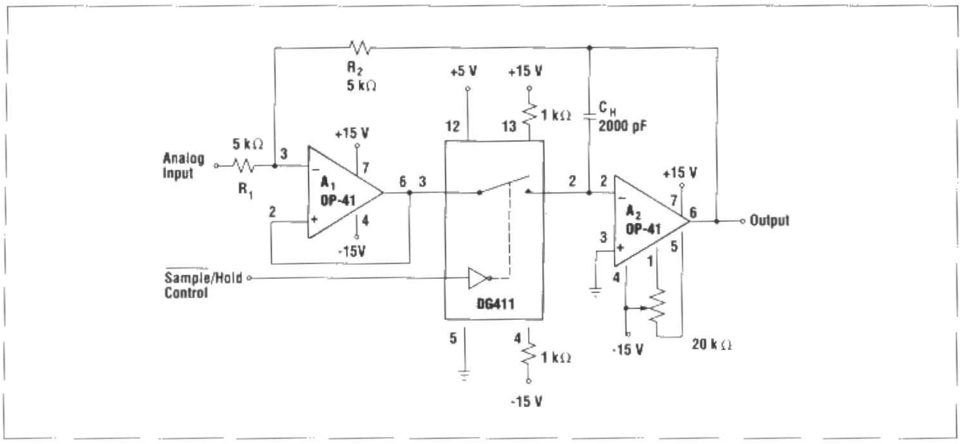



Fig. 5. Integrator output sample-and-hold function operates switch into virtual earth.

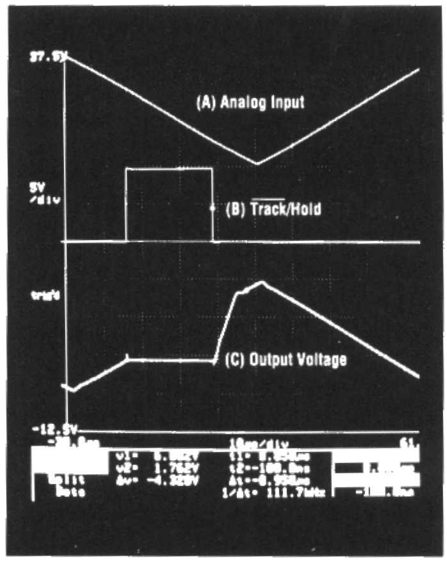



Fig. 6. Acquisition time is limited by the slew rate of the output amplifier.

remains after the upset and the droop rate that begins after settling is completed.

## Closed-loop integrator output sample-and-hold circuit

A popular sample-and-hold configuration is shown in Fig. 5. This circuit is simple and accurate. It has a gain of –1 since $R_1 = R_2$. Opamp A1 acts as a current booster to speed up the charging rate of hold capacitor $C_H$. Since the unity-gain buffer has a very low output impedance, the time constant associated with $C_H$ is determined primarily by the on-resistance of the switch and by the magnitude of the hold capacitor. Thus, the circuit benefits greatly from the low on-resistance of the DG411.

The settling time of the output voltage is determined by the slew rate and settling time of the integrator stage. In the sample mode of operation, the DG411 closes and hold capacitor $C_H$ charges to the negative of the input voltage. In the hold mode, the analogue switch opens after the capacitor has acquired this voltage to the desired accuracy. Another advantage of the DG411 is that the switch always operates at a virtual earth potential regardless of the input voltage. Since at this level the charge injection on the switch drain is at its minimum value, the hold step error is minimized. The errors of A1 are minimized in the sample state, although they do appear in the hold mode.

The photograph in Fig. 6 shows the typical waveforms associated with this circuit. With the components shown, this circuit achieved an acquisition time of about 20 µs, a maximum hold step error of 3.8 mV and a droop rate of 7.5 µV/µs.

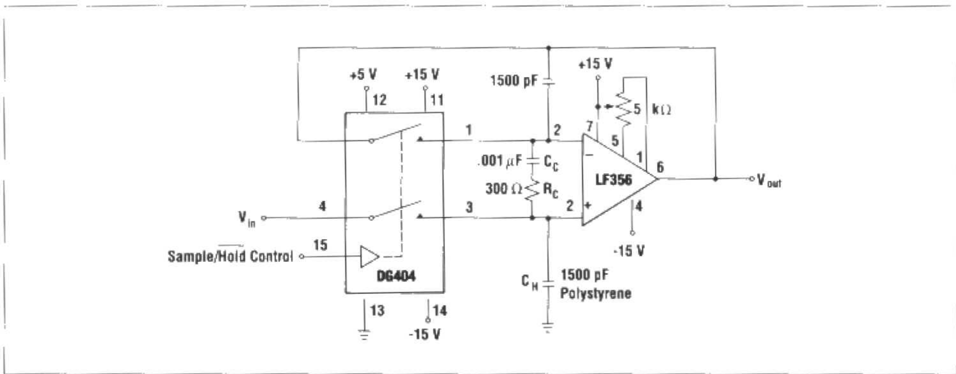



Fig. 7. Fast and precise sample-and-hold circuit.

## Fast and precise sample-and-hold circuit

The circuit shown in Fig. 7 uses a DG404 analogue switch in conjunction with a JFET input operational amplifier. The DG404 is a fast switch ($T_{on}$<150 ns). In this circuit, both switches have a similar potential when open, so their charge injkection effect is minimized by their differential effect on the opamp. Acquisition time of this circuit was less than 600 ns, worst-case pedestal error was –5 mV, and droop rate was 35 µV/µs.

The compensation network formed by $C_C$ and $R_C$ helps to reduce the hold-time glitch and optimizes acquisition time. The photograph in Fig. 8 shows this circuit's output without a compensation network. Notice the large glitch going into the hold mode, as well as the rippled waveform right after the output slews to its new value at settling time. The photograph in Fig. 9 shows the improved response after the compensating network has been installed.

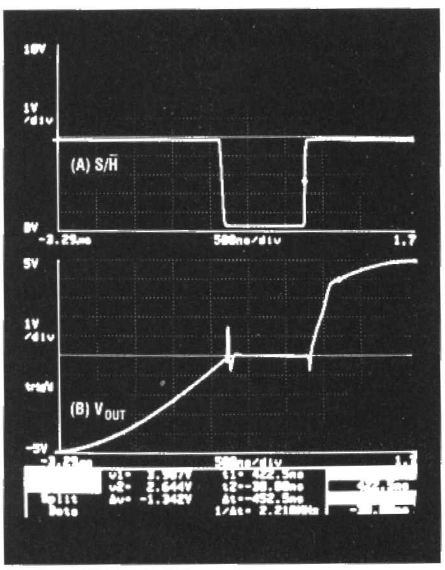



Fig. 8. $V_{out}$ without compensation shows large glitches and a waveform ripple during acquisition time.

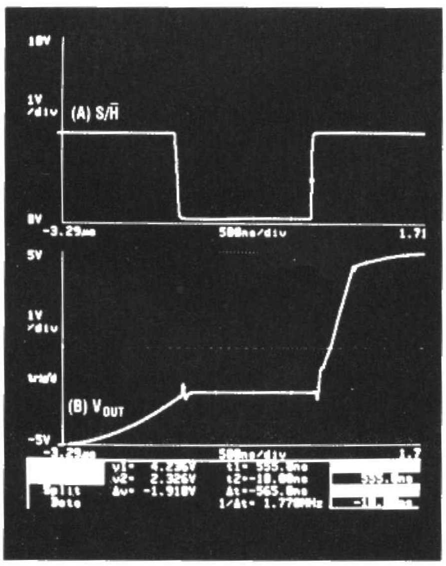



Fig. 9. Improved $V_{out}$ after compensation.

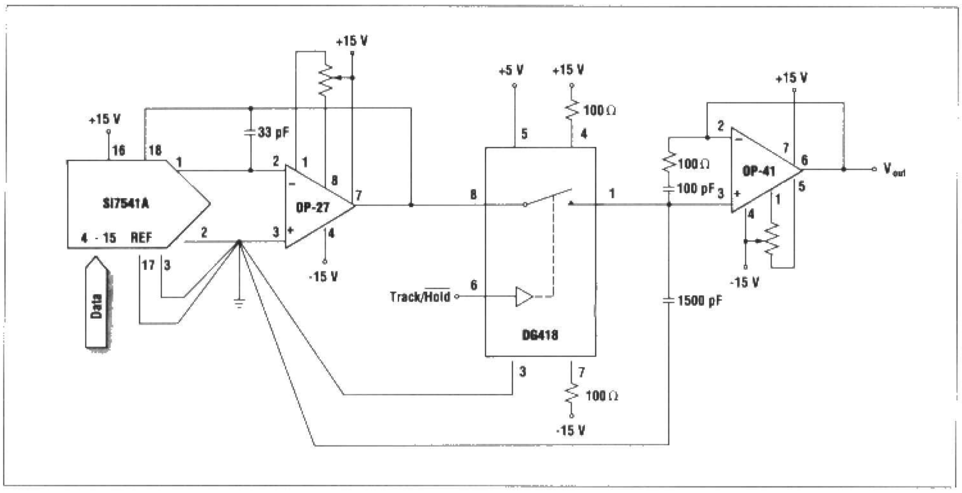



Fig. 10. Digital-to-analogue converter deglitcher.

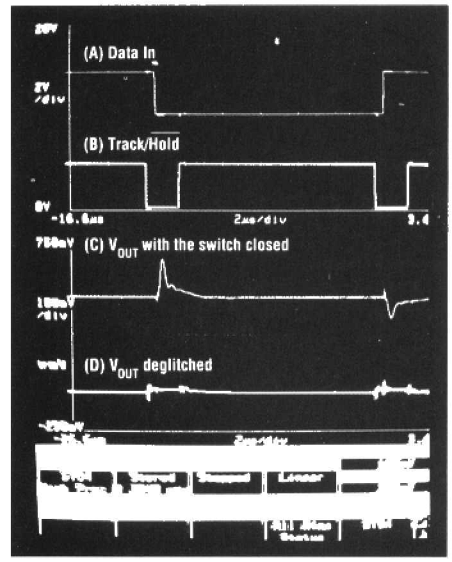



Fig. 11. Digital-to-analogue converter deglitcher waveforms.

## Digital-to-analogue converter deglitcher

Major code transitions in digital-to-analogue converters (DACs) can cause unwanted voltage spikes, commonly called glitches. In many DAC applications, these glitches can not be tolerated. Additionally, DACs from different vendors have different size glitches. (Note the glitch impulse specification on DAC data sheets). To ensure a smooth transition when the DAC goes from one voltage to the next and to guarantee uniform circuit response regardless of alternate-sourced DACs, the DAC output may be processed with a tack-and-hold as shown in Fig. 10. While the DAC input code is unchanged, the DG418 is closed and $V_{out}$ tracks the output of the current-to-voltage converter. Just before a code change occurs, the analogue switch is opened so that $V_{out}$ continues showing the previous voltage. After the code change and its associated glitch has settled, the DG418 closes again and the track mode is resumed.

The photograph in Fig. 11 shows $V_{out}$ with the DG418 always closed (c) and with the deglitcher active (d). Notice the improvement in the transition glitches.

The DG418 offers high switching speeds, which are required for short conversion times, and low charge injection, which minimizes pedestal errors.

## Dual-input programmable gain amplifier

For digital systems where only a +5 V supply is available, a small amount of analogue processing can be implemented with a low-voltage converter IC and low-voltage analogue components. Figure 12 shows an amplifier suitable for data acquisition or voice recognition applications where either of two analogue signals is selected and amplified by a very precise, self-calibrating chopper-stabilized amplifier. Circuit gain can be selected as either ×2 or ×10. A single DG423 analogue switch was used to perform both the input-select and gain-select functions. Its low on-resistance, high speed, and on-chip latches ease circuit design and improve overall accuracy.

The photograph in Fig. 13 illustrates the operation of the circuit. For demonstration purposes, input- and gain-selects were tied together so that when the 0.5 V (p-p) triangular signal was being processed, the circuit gain was ×10 whereas when the 3 V (p-p) sine wave was selected, the amplifier's gain was reduced to ×2. This type of gain ranging is useful to precondition analogue signals of different amplitudes prior to an analogue-to-digital conversion.

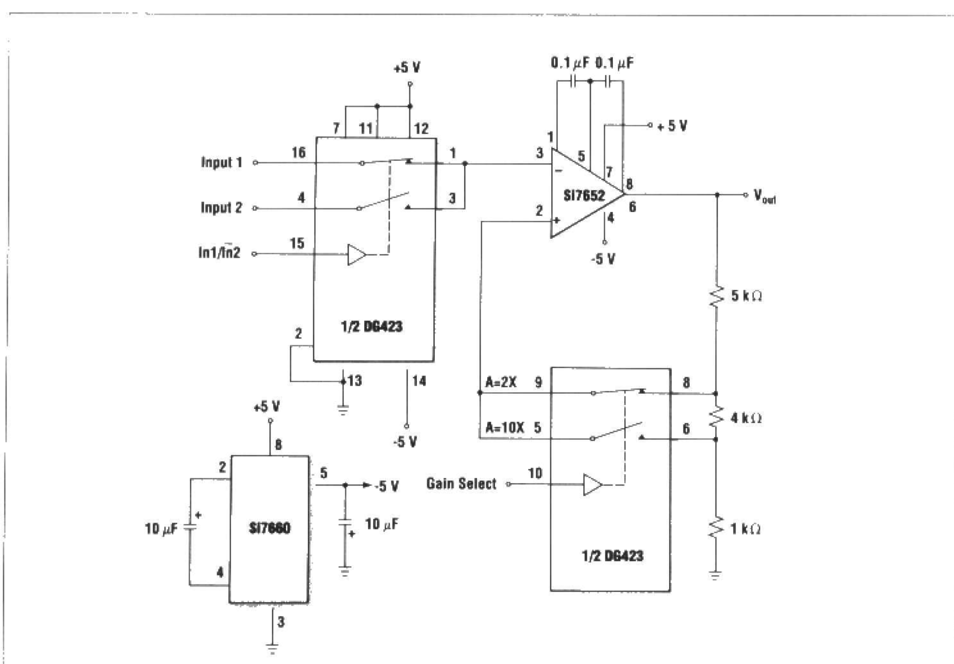



Fig. 12. Low-voltage programmable gain amplifier.

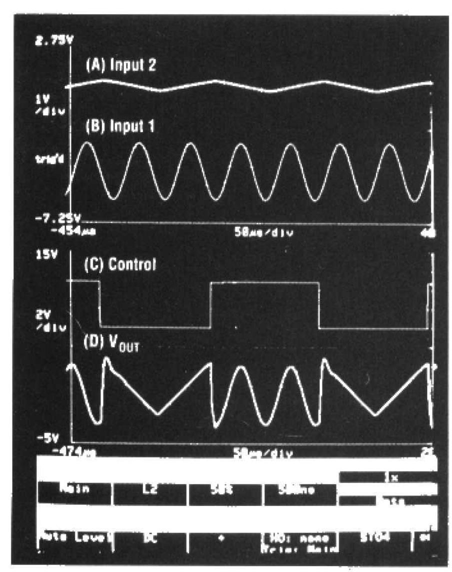



Fig. 13. Gain ranging produces similar amplitudes even if the input levels are different.

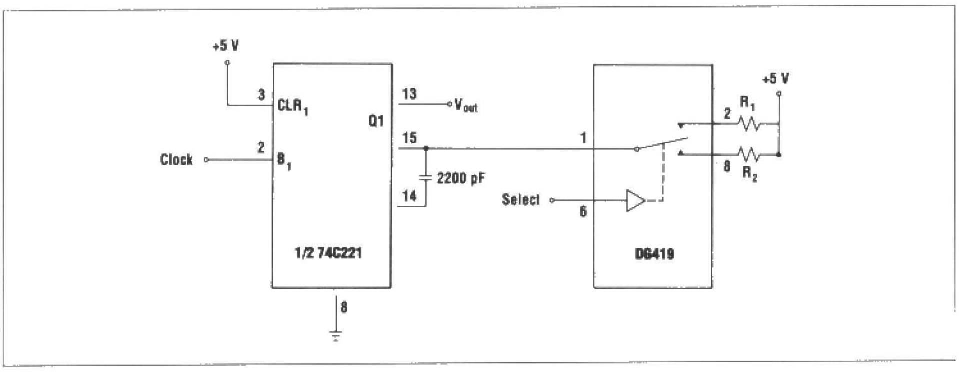



Fig. 14. Programmable one-shot multivibrator.

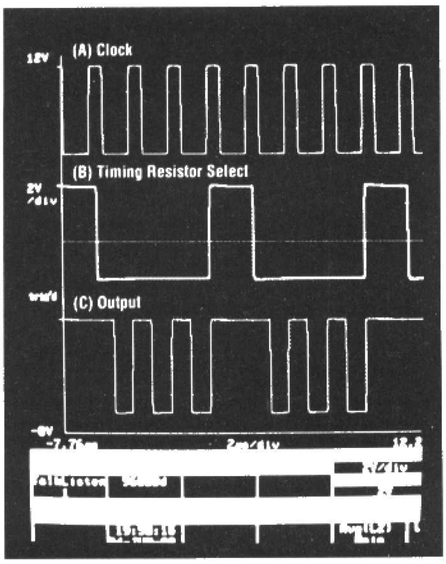



Fig. 15.. A logic LOW produces short pulses and a logic HIGH creates long ones.

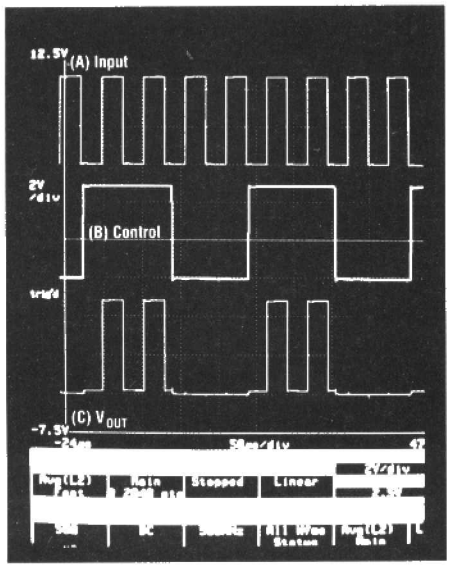



Fig. 16. This photograph illustrates the remote switch-over action.

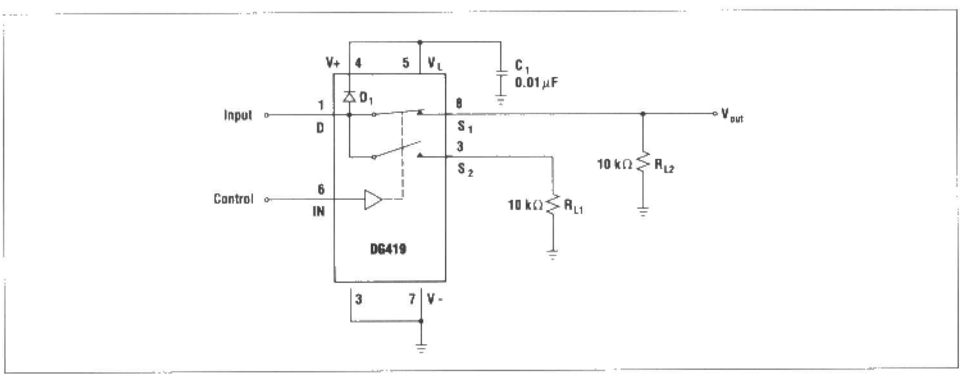



Fig. 17. Remote SPDT analogue switch for switched signal powers.

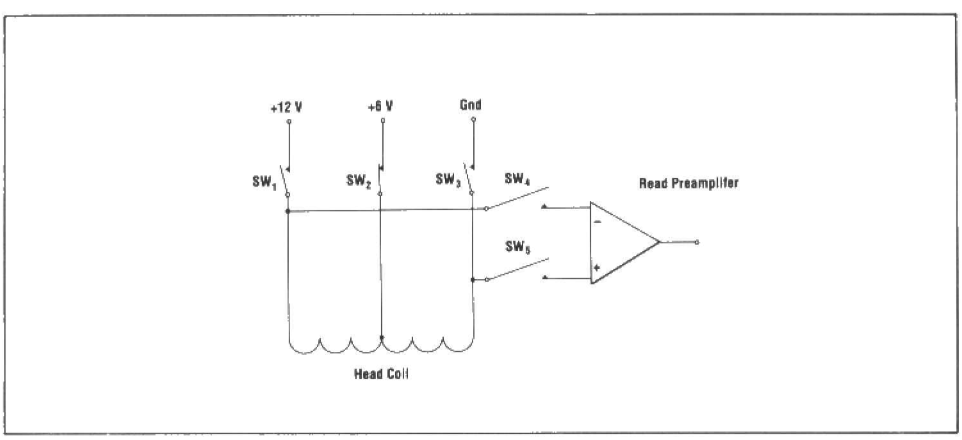



Fig. 18. DG411s in the head switching circuit of a disk drive.

## Programmable one-shot multivibrator

Another useful application for an analogue switch, a programmable one-shot multivibrator, is shown in Fig. 14. This circuit produces pulses whose duration is determined by digitally selecting one of the two timing resistors—see Fig.15. Advantages of the use of the DG419 in this circuit are: small size (8-pin miniDIP or small-outline package), high speed, low on-resistance, and TTL compatibility even in single-supply operation.

## Analogue switch powered by input signal

The analogue switch in Fig. 17 derives operating power from its input signal, provided that the amplitude of that signal exceeds 4 V and the frequency is greater than 1 kHz. This circuit is useful when signals are to be routed to either of two remote loads. Only three conductors are required: one for the signal to be switched, one for the control signal and a common return.

A positive input pulse – see Fig. 16 – turns on clamping diode $D_1$ and charges $C_1$. The charge stored on the capacitor is used to power the chip; operation is satisfactory because the switch requires a supply current of not greater than 1 μA. Loading of the signal source is imperceptible. The DG419's on-resistance has the respectable value of 100 Ω for an input signal of 5 V.

## Read/write disk-drive circuit

The circuit shown in Fig. 18 allows data to be written to or read from a disk. In the write mode, SW2 is closed. A ONE is created by momentarily closing SW1. This causes current to flow in the left-hand half of the head coil. A ZERO is produced when SW3 is closed. This causes current to flow in the right-hand half of the coil and reverses the direction of the magnetic flux.

In the read mode, switches SW4 and SW5 are closed. This connects the head coil to the read preamplifier so that the voltages picked up by the head as the disk glides by can be amplified.

Single-supply operation with +12 V, low-on resistance and high switching speed allow an improvement in data rates of roughly ×10 when DG411s are used in place of the more mature DG211s.

## Micropower UPS transfer switch

The purpose of the uninterrupted power supply (UPS) circuit in Fig. 19 is to preserve volatile memory contents in the

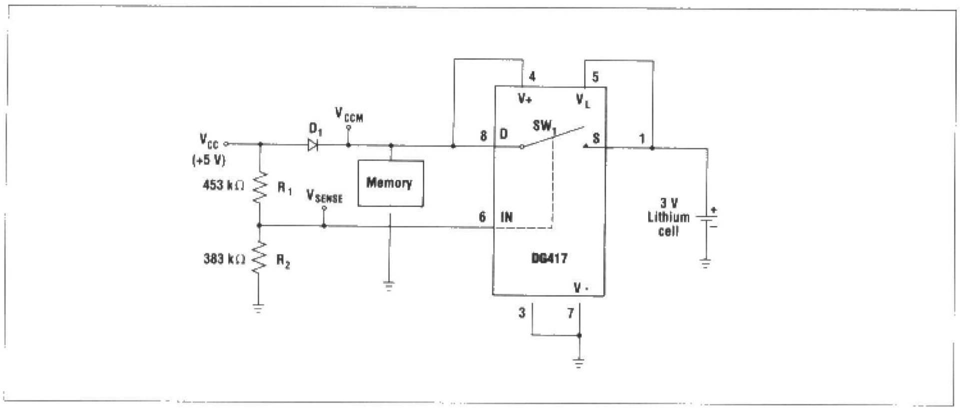



Fig. 19. Micropower UPS circuit.

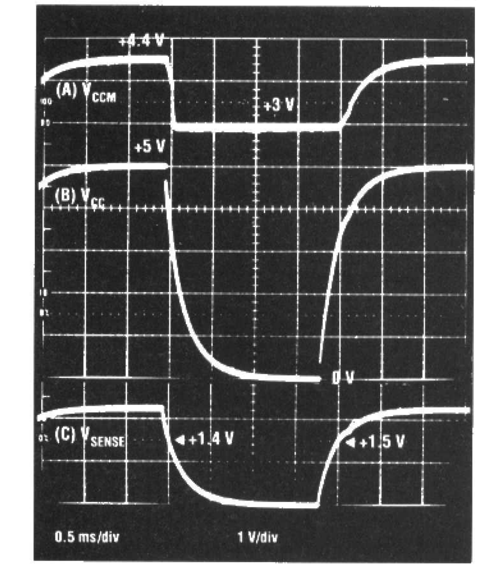



Fig. 20. Oscilloscope waveforms show a clean power switch-over.

event of a power failure. In this application, every tenth of a volt counts. This circuit uses a micropower analogue switch that comes in an 8-pin miniDIP or small-outline package, a 3-V lithium cell to supply back-up power, a diode and two resistors. Voltage losses under 0.1 V can be achie-ved.

During normal operation, currents of several hundreds milliamperes are supplied from $V_{CC}$. In this mode, SW1 is open, so that the only drain from the lithium cell consists of leakage currents flowing into the $V_1$ and S terminals. The leakage current is typically about 10 pA. Resistors $R_1$ and $R_2$ are continuously sampling $V_{CC}$.

When $V_{CC}$ drops to 3.3 V, the DG417 changes states, closing SW1 and connecting the back-up cell. Diode $D_1$ prevents current from leaking back towards the rest of the circuit. Current consumption by the CMOS analogue switch is around 100 pA: this ensures that most of the available power is applied to the memory where it is really needed. In the stand-by mode, currents of some hundreds of milliamperes are sufficient to retain data.

When the +5 V supply comes back on, the potential divider senses the presence of at least 3.5 V and causes a new change of state in the analogue switch, restoring normal operation.

On-resistance is about 74 Ω when $V_{CC}$ is +5 V and 128 Ω when $V_{CC}$ is +3 V. For example, an 800 μA load, equivalent to a static RAM of 256 kbit (MCM61L16), will produce a voltage drop of 0.1 V on the analogue switch, which is much better than the 0.6 V drop occurring if a simple 2-diode circuit were used.

Higher currents and lower losses can be achieved by paralleling several sections in a multiple analogue switch such as the DG403.

Line a in the photograph in Fig. 20 illustrates how, in spite of $V_{CC}$ dropping to 0 V (line b), uninterrupted power is applied to the load. Negligible voltage loss is caused by the switch. Line c shows that the DG417 changes state when its control input voltage decays to 1.4 V and changes again when it reaches 1.5 V on its way back to normal. The values of $R_1$ and $R_2$ may be adjusted for different trip points if desired.

For the applications mentioned in this article, the DG4090 family of silicon-gate CMOS switches comes a step closer to the ideal switch. Any application that uses industry-standard analogue switches can now be improved by choosing these fast, lower-power, versatile analogue switches.

# ELECTRONICS SCENE

### Course for the Radio Amateurs' Exam – May 1990

There will be a course for the May 1990 Radio Amateurs' Exam at Newark Technical College, Chauntry Park, Newark, Notts, starting this month on Monday evenings from 7 to 9 p.m. The course tutor is Alistair Morrison G4YZG.

Further details may be obtained from Bert Drury G1UMK at the college, telephone (0636) 705921.

### Radio Amateurs' Course

North Trafford College, Manchester is again offering a Radio Amateurs' Course starting this month. The course will comprise:

| | |
|---|---|
| Theory | Thursday evening or Wednesday morning. |
| Morse Code | Tuesday evening or Wednesday afternoon |
| Amateur TV | Wednesday morning |
| Advanced Morse Code | Monday evening |

Lecturer: J.T. Beaumont G3NGD.

Enrolment dates are 6, 7 and 8 September. Further details from North Trafford College of Further Education • Talbot Road • Stretford • Manchester M32 0XH • Telephone 061 872 3731.

### New Showroom for Nevada

Nevada, the specialist suppliers of Communications, Music and Discotheque equipment have opened a new showroom in Portsmouth, adjacent to their present premises at 189 London Road, North End, Portsmouth PO22 9AE, telephone (0705) 662145.

In the new showroom, radio, scanner and short-wave enthusiasts will be able to browse in comfort with full 'hands on' facilities over Nevada's expanded range of products.

### New Maplin Store

Maplin Electronics, one of the fastest growing electronic retail groups in Europe, has recently opened another store in London.

The new store, managed by Lawrence Saunders, is located at 146-148 Burnt Oak Broadway, Edgware, Middlesex.

Further details from Maplin Head Office, telephone (0702) 552911.

# THE DIGITAL MODEL TRAIN – PART 6

by T. Wigmore

**The sixth part in the series deals in detail with the Booster Unit. Each of these amplifiers provides enough power for the control of up to fifteen trains on a digital model track. The booster is the last unit in the series that can be used with both the Märklin and the Elektor ElectronicsDigital Train System. The units featured in forthcoming parts in the series are peculiar to the Elektor Electronics System.**

The power supply of a digitally controlled model railway track is fundamentally different from that of a conventional track, in that the supply voltage is switched rapidly between +18 V and −18 V. The switching is carried out by the booster (power amplifier).

The booster ensures that the serial control commands generated by the digital control circuits contain not only the information, but also the power to start locomotives, turnouts (points) and signals.

Since derailments, and the consequent short-circuits of the track, occur much more often on model railway tracks than on life-size ones, it is essential that the booster is provided with an efficient short-circuit protection facility.

## The concept

Our booster unit has two important advantages over that from Märklin: higher output power and a regulated output voltage.

The Märklin booster provides a maximum output current of about 3 A. That is not much if you take the current drawn by one locomotive at about 700 mA , and add to this the current drawn by turnouts (points), signals, and coach lighting. It is on those considerations that our booster provides an output current of 10 A.

The output voltage of the Märklin booster is fairly load-dependent: a 25% drop over the normal range of loads is quite normal. That kind of variation has, of course, an adverse effect on the speed of the locomotives and the brightness of the coach lights.

The output stage is an emitter follower. Driving the bases by a voltage source ensures a virtually constant output voltage, which results in independent speed control of the trains and constant brightness of the various lights. These properties are illustrated in Fig. 39 and Fig. 41.

The use of an emitter follower also enables higher switching speeds since the transistors operate on the linear part of their characteristics: the switching times are, therefore, not adversely affected by saturation effects.

A drawback of the configuration is the higher voltage and the consequent greater dissipation in the output transistors. Fortunately, this is easily rectified by the use of somewhat larger heat sinks.

## The circuit

Since the switching pattern of the track voltage contains control information, it is important that the booster provides a clean output signal. Much attention has, therefore, been paid to the switching speed. The practical outcome is illustrated in Fig. 40.

The bases of the emitter follower, $T_1$–$T_4$ in Fig. 42, are switched by $T_5$ and $T_6$ respectively between +20 V and −20 V. These voltages are provided by $IC_1$–$D_3$ and $IC_2$–$D_4$ respectively. The final output voltage is the difference between the base voltage and the sum of the base-emitter potential (about 1.5 V) of the output transistors and the drop across the emitter resistors (maximum 0.6 V). In practice, the output voltage is a reasonably constant ±18 V. See also the load characteristic in Fig. 41.

The emitter follower ensures a better bandwidth and regulation with complex loads than, for instance, feedback.

Emitter resistors $R_{12}$–$R_{15}$ ensure an equal division of current to $T_1$–$T_2$ on the one hand and to $T_3$–$T_4$ on the other.

Resistors $R_{12}$ and $R_{14}$ serve to measure the current in aid of short-circuit protection transistors $T_9$ and $T_{10}$. When the emitter current of $T_1$ or $T_3$ tends to become too high, the drop across $R_{12}$ or $R_{14}$ rises sufficiently to switch on $T_9$ or $T_{10}$. This causes a reduction in the base current of the output transistors and, consequently, in their collector and emitter currents.

The input stage is formed by $T_7$ and $T_8$ and is configured in a manner that makes a symmetrical input signal essential. If the input (pin 4 of $K_1$) is 0 V or not connected,

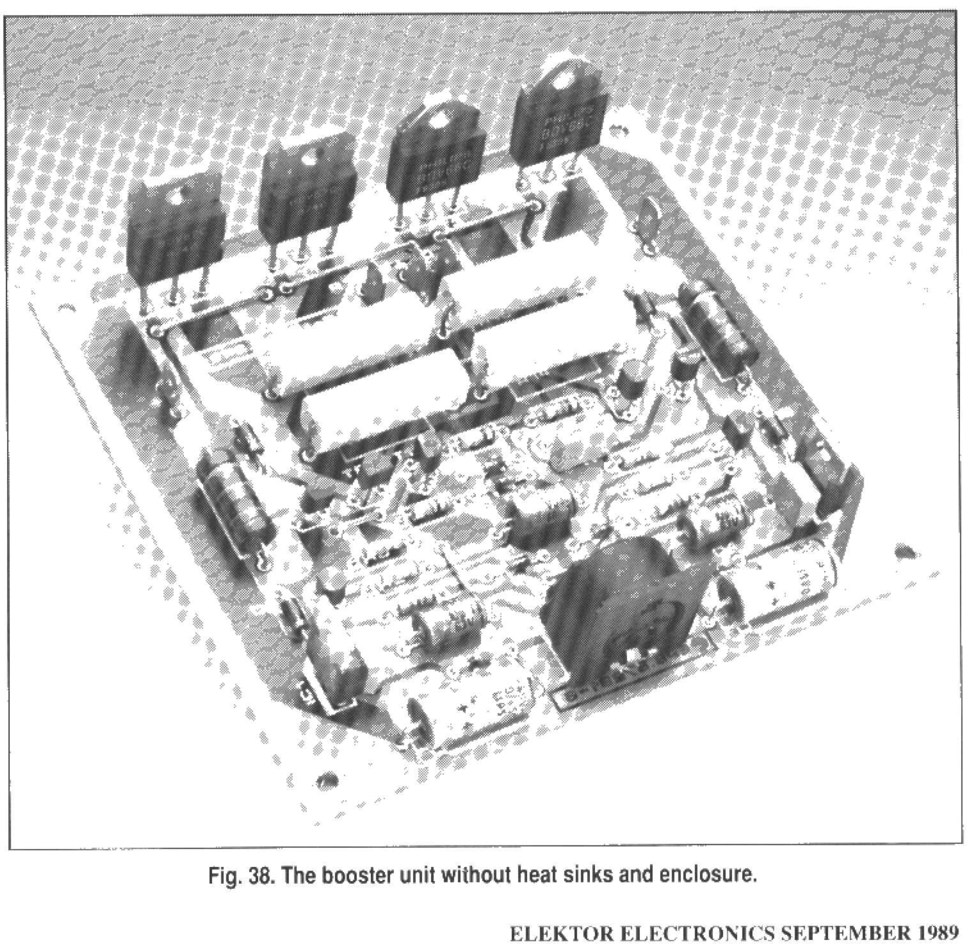


Fig. 38. The booster unit without heat sinks and enclosure.

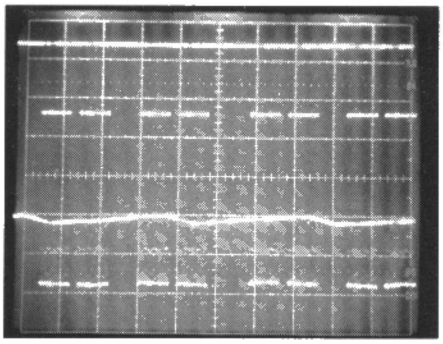


Fig. 39. Data transfer by switching the supply voltage. It is evident that the Märklin Booster (upper trace) does not provide a regulated output in contrast to the Elektor Electronics unit (lower trace).

all transistors are switched off and the output presents a high impedance (that is, no voltage is supplied to the rails). When the input voltage is between +5 V and +20 V, T7, T5, T1, and T2 conduct and the output is switched to +18 V. With the input voltage between −5 V and −20 V, T8, T6, T3 and T4 conduct and the output voltage is switched to −18 V.

All transistors, except T5 and T6, operate on the linear part of their characteristics. Transistors T5 and T6 are switched in the saturation region because switching transistors for voltages of 50 V and more are not available. Nevertheless, C8 ensures that these transistors switch at a sufficiently high speed.

## Overload signal

The circuit around T11 serves to indicate an overload condition. Note that only the negative output voltage is monitored. This is sufficient since the load on the negative line is slightly higher than that on the positive rail. For instance, the turnout (points) decoders work with half-wave rectifiers and, therefore, load only the negative rail. Moreover, when no data are being transmitted, the output voltage is negative.

When the booster is overloaded, T9 and T10 limit the current in the first instance. The output voltage will then drop significantly and this causes a rise in the voltage across the output transistors and thus in the dissipation. If this situation is allowed to persist, there is a danger of the booster being thermally overloaded: the consequent risk of fire is a very real one.

Therefore, if the output voltage drops below 15 V, T11 will switch off. The signal at pin 5 of K1, aided by the pull-up resistor on the main PCB of the Elektor Electronics system, goes low and this results in the removal of the drive to the booster with the aid of the software. Thermal overloads are, therefore, prevented; moreover, the system 'knows' that in this condition no data can be transmitted (even if they could, they would not reach the decoders).

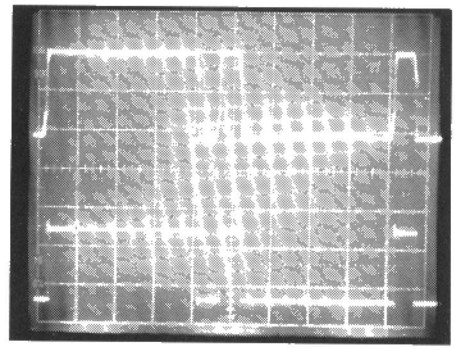


Fig. 40. Comparison of the switching behaviour under load of the Märklin Booster (upper traces) and the Elektor Electronics unit (lower traces).

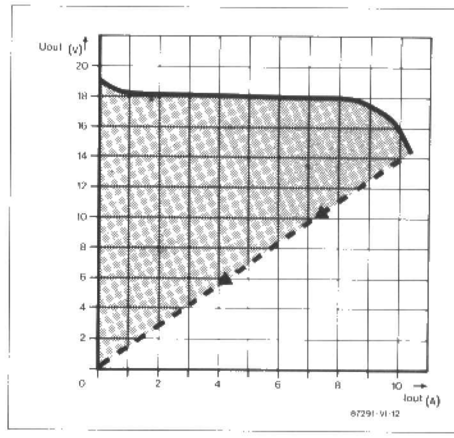



Fig. 41. Load characteristic of the booster unit.

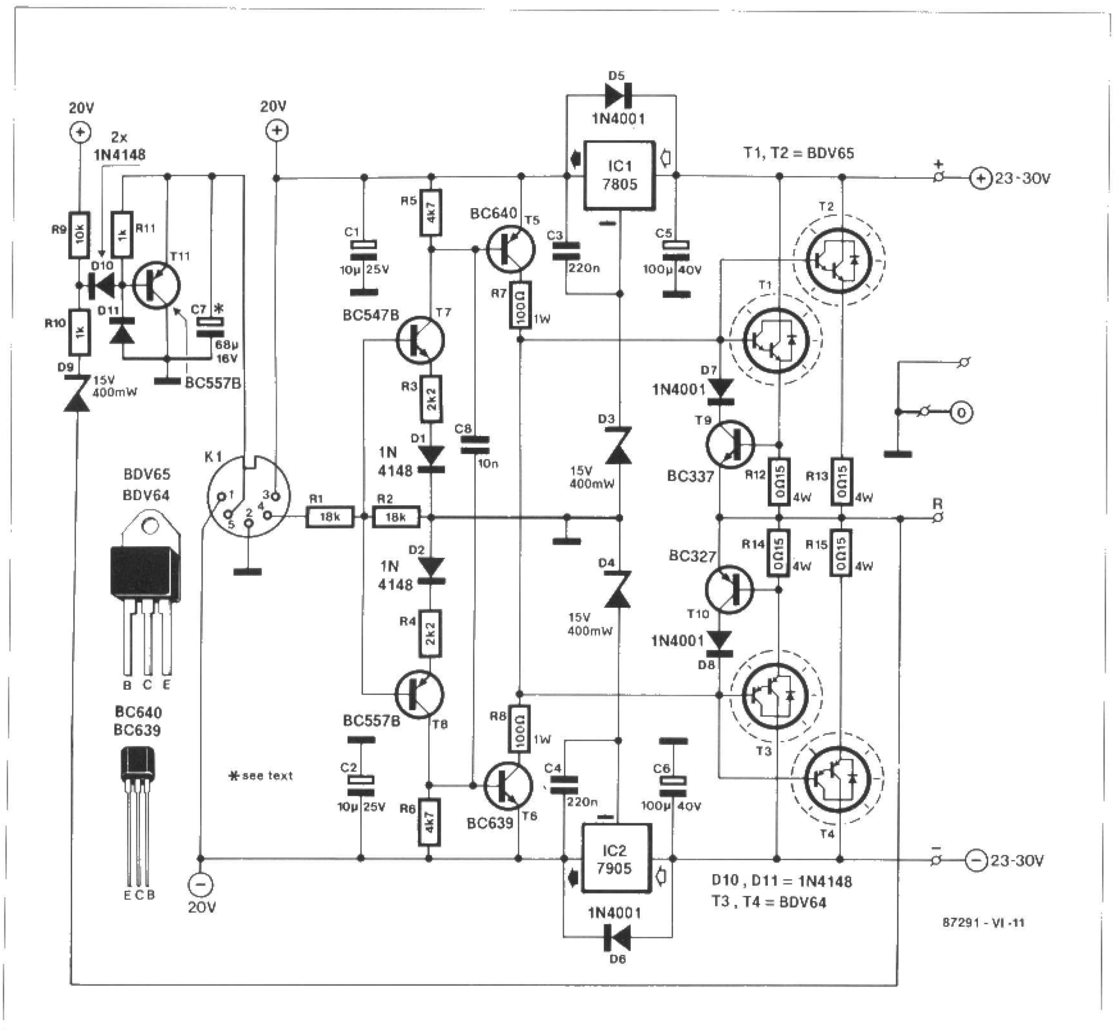



Fig. 42. Circuit diagram of the booster unit.

Capacitor C7 enables the overload action to be delayed, so that the system is not disabled at every momentary short circuit. This will be reverted to later in the series.

## Construction

If the PCB shown in Fig. 45 is used, construction of the booster unit should not present any problems.

Fit the wire links first: those close to the output transistors should be of 1 mm dia. wire.

Mount resistors R12–R15 well away from the board, because they get pretty hot during operation.

The board has provision for a 5-pin DIN connector, but if the booster is intended for use in a stationary position (which is normally the case), the respective wires may be soldered direct to the board.

Circuits IC1 and IC2 do not need a heat sink.

Do not fit C7 at this stage.

Transistors T1–T4 must be mounted on a heat sink with a thermal resistance of not less than 0.8 K/W with the aid of good-quality insulating

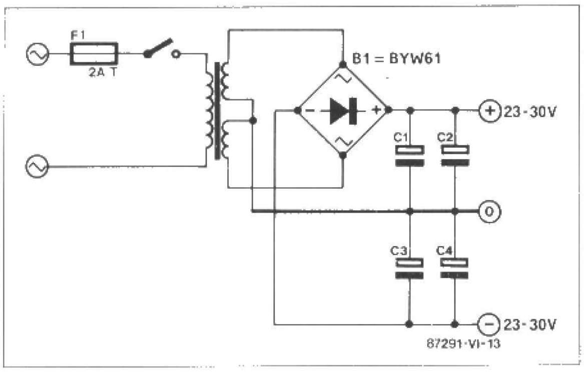



Fig. 43. Circuit diagram of a recommended power supply.

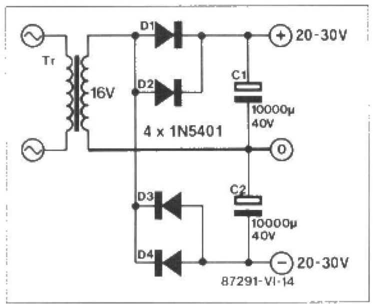



Fig. 44. If you are content (for the time being) with a much smaller output current, this power supply will do nicely.

washers. If BDX66/67 darlingtons (with TO-3 housing) are used, they must first be mounted on to the heat sink and then connected to the board by not too long, heavy-duty wires.

## Power supply

The circuit diagram of a recommended power supply is shown in Fig. 43. The 2×18 V transformer must preferably be a toroidal type. The rectifier must be a heavy-duty type and needs a heat sink (it may be mounted on to that for $T_1$–$T_4$).

If you do not want to go to the expense of a new transformer, but rather use one that you have had lying around for some time, use the circuit shown in Fig. 44. Remember, however, that such a set-up will normally not be able to deliver more than a quarter of the power of the supply in Fig. 43, if that.

Finally, DO NOT connect transformers in parallel to increase the total available current: such a set-up can be a death trap.

## Assembly and test

Since the booster is operated from the mains, great care and attention must be paid to correct assembly and insulation. Because there are always metal parts in a model railway system that can be touched (like the rails), it is advisable to use a good-quality insulated enclosure.

The insulation of the power supply transformer stated in the parts list is approved to Class I. This means that the mains cable should have three cores, one of which is earth.

All metal parts that can be touched (including the heat sinks) should be connected to earth.

Connect the two secondary windings of the mains transformer in Fig. 43 in series and fit and solder the rectifier and the buffer capacitors ($C_{tot} = C_1 + C_2 = C_3 + C_4 = \geq 20{,}000\ \mu F$ rated at $\geq 40$ V).

Before the supply is connected to the booster, switch on the mains and check that the direct voltage across the buffer capacitors is ±25–29 V. If you measure 0 V, it is almost certain that the two secondary windings have been connected in anti-phase. Switch off the mains, discharge the capacitors via a resistor and reverse the connections of one of the secondary windings. Then check the direct voltage again.

If everything is all right, switch off the mains again and discharge the buffer capacitors via a resistor.

Next, connect the supply to the booster via insulated wire of at least 0.5 mm dia. and switch on the mains. Check that the

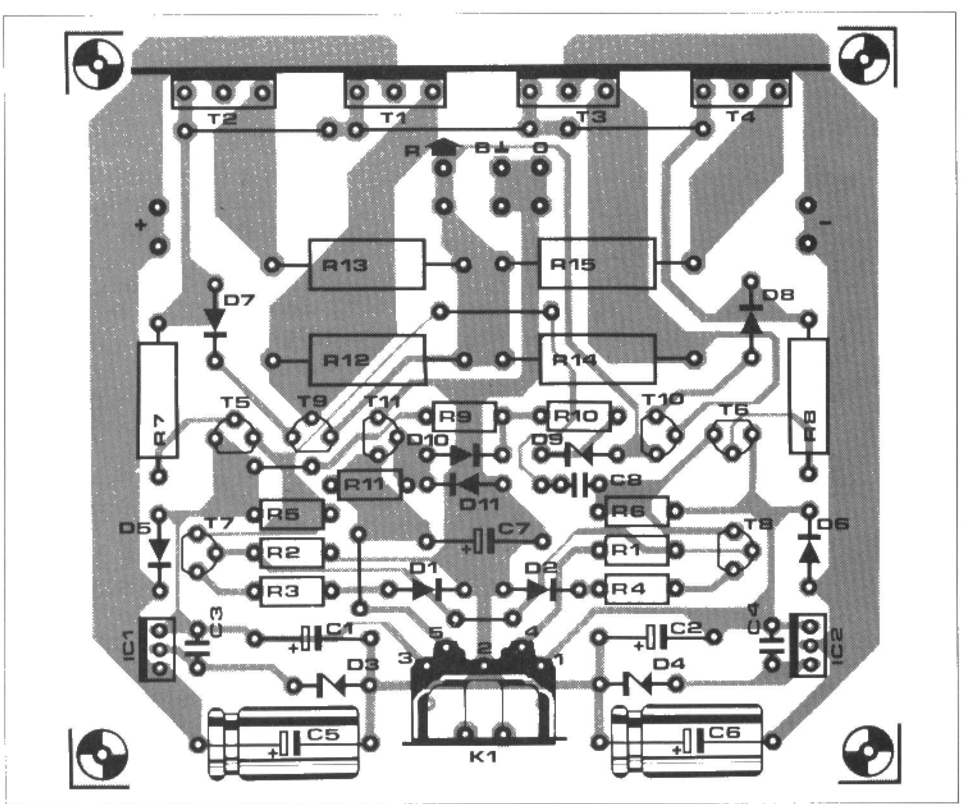



Fig. 45. Printed circuit board for the booster unit.

**Parts list**

**Resistors:**
$R_1$;$R_2$ = 18 k
$R_3$;$R_4$ = 2k2
$R_5$;$R_6$ = 4k7
$R_7$;$R_8$ = 100Ω; 1 W
$R_9$ = 10k
$R_{10}$;$R_{11}$ = 1k0
$R_{12}$(196$R_{15}$ = 0Ω15; 4 W

**Capacitors:**
$C_1$;$C_2$ = 10μ; 25 V
$C_3$;$C_4$ = 220n
$C_5$;$C_6$ = 100μ; 40 V
$C_7$ = 68μ; 16 V
$C_8$ = 10n

**Semiconductors:**
$D_1$;$D_2$;$D_{10}$;$D_{11}$ = 1N4148
$D_3$;$D_4$;$D_9$ = zener diode 15 V; 400 mW
$D_5$–$D_8$ = 1N4001
$T_1$;$T_2$ = BDV 65 (Philips Components)
$T_3$;$T_4$ = BDV 64 (Philips Components)
$T_5$ = BC640
$T_6$ = BC639
$T_7$ = BC547B
$T_8$ = BC557B
$T_9$ = BC337
$T_{10}$ = BC327
$T_{11}$ = BC557
$IC_1$ = 7805
$IC_2$ = 7905

**Miscellaneous:**
$K_1$ = 5-way DIN socket (180°) for PCB mounting.
5 off car-type spade terminals for PCB mounting.
Insulation material for $T_1$–$T_4$.
PCB Type 87291-6 (see Readers Services page).

*Recommended power supply parts (not on PCB):*
Mains transformer: 2×18 V @300 VA (e.g. ILP 73014)
Smoothing capacitors: 4 off 10,000μF; 40 V or 4 off 15,000μF; 40 V.
High-current bridge rectifier: min. 20 A (e.g., BYW61 from Motorola).
One mains-rated fuse: 2 A slow.
Two low-voltage fuses: 10 A fast .
Heat-sink: e.g., SK120-100mm (Dau Components; Fischer).

output voltage of IC1 is +20 V and that of IC2 is −20 V. There should be no voltage between B (earth) and R since there is as yet no input.

Fit a 100 Ω, 5 W, resistor between B and R and connect the input, pin 4 of K1, to a positive voltage, for instance, +20 V at pin 3 of K1. The output voltage should then be +18 V. With a negative input – obtained by interconnecting pins 1 and 4 of K1 – the output should be −18 V.

## Connecting to Märklin Digital

The booster circuit is driven via K1, which also carries the auxiliary +20 V and −20 V voltages. These are not of importance when the Märklin Digital is used, but in the Elektor Electronics system they power the RS232 interface.

When the Märklin Digital is used, only pins 2 (earth) and 4 (input) of K1 need to be connected to the brown and red terminal at the rear of the Central Unit. Our booster, therefore, does NOT use the 5-pin connector on the Central Unit.

The overload signal (pin 5 of K1) is also for use with our own system only. To arrange for the automatic switch-off of the Märklin unit during overloads, a diode must be added as shown in Fig. 46. Warning of a short circuit in the booster is then passed to the Central Unit, after which the current monitor in that unit arranges the switch-off.

The Central Unit may provide some of the power to the rails, but note that only the B connexions of the Central Unit and our booster may be interlinked. The R connexions (centre rail in the Märklin system) must be isolated from one another. Märklin supplies special parts to prevent the slide contacts from short-circuiting the electrically separated centre rails during cross-overs.

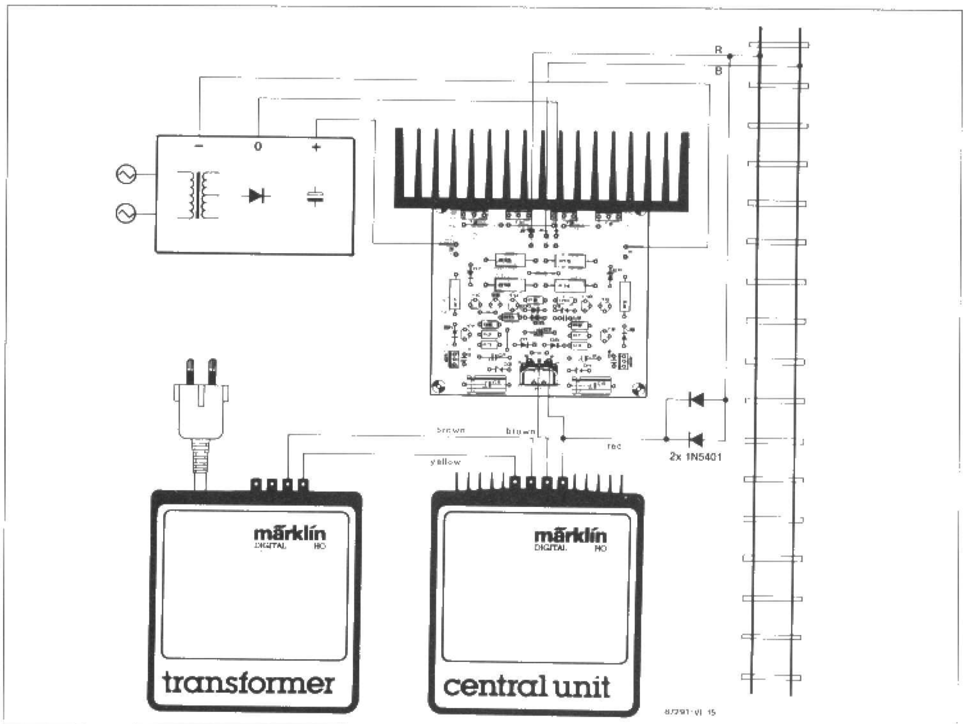



Fig. 46. How to connect the booster unit to Märklin's Central Unit.

## For true-to-scale modellers

The output voltage of the booster was chosen at ±18 V to ensure that the maximum speed of the locomotives would be about equal to that in traditional model railway systems. Taken to scale, model trains travel faster than life-size ones. Modellers who want to have their locomotives travel at, proportionally, the same speed as life-size trains can arrange this by using 12 V zener diodes in the D3, D4 and D9 positions. This will result in an output voltage of ±15 V.

# EVENTS

**IEE Meetings**

3–8 Sep — Personal and mobile radio systems (Vacation School at Swansea).
4–7 Sep — Microwave (Conference at Wembley Conference Centre).
5–8 Sep — Circuit theory and design (Conference at Univ. of Sussex).
17–22 Sep — The application of artificial intelligence (Vacation School at Univ. of Reading).
17–22 Sep — Data communications and networks (Vacation School at Aston University).
17–22 Sep — Integrated electronic product design (Vacation School at University of Surrey).
18–20 Sep — Software engineering for real-time systems (Conference at Cirencester).
18–22 Sep — Quantum electronics (Conference at University of Oxford).
24–30 Sep — Microwave measurement (Vacation School at Univ. of Kent).

Further information on these, and many other, events from IEE • Savoy Place • LONDON WC2R 0BL • Telephone 01-240 1871.

The **Third International Symposium on IC Design and Manufacture** will be held on 14–15 September in Singapore. Details from the School of Electrical and Electronic Engineering • Nanyang Technological Institute • Nanyang Avenue • SINGAPORE • Telephone Singapore 2263.

A series of short courses on **Graphics and Image Processing, Microprocessors, Signal Processing, Supercomputers, Computer Networks, Voice & Data Communications,** and many others, will be conducted by ICS • Trafalgar House • Hammersmith International Centre • LONDON W6 8DN • Phone 01-748 6667.

A seminar on **Mid-range systems and UNIX** will be held at the QEII Centre, Westminster, London on 27–28 September. Details from Blenheim Online • Blenheim House • Ashill Drive • PINNER HA5 2AE • Telephone 01-868 4466.

A number of seminars has been organized for this month by **Frost & Sullivan.** Subjects include: **Information Technology; Telecommunications & Data Communications; and Electronic Engineering.** Details from Frost & Sullivan • Sullivan House • 4 Grosvenor Gardens • LONDON SW1W 0DH • Telephone 01-730 3438.

Synopses of papers for the **Internation al Conference on Electromagnetic Compatibility** (University of York: 28–31 August 1990) should be submitted before 29 September; those for the **Fourth International Conference on Advanced Infrared Detectors and Systems** (London:5–7 June 1990) before 29 August; and those for the **International Conference on Integrated Broadband Services and Networks** (London: 15–18 October 1990) before 15 January to: Conference Services • IEE • Savoy Place • LONDON WC2R 0BL • Telephone 01-240 1871 Ext. 222.

# INTERMEDIATE PROJECT

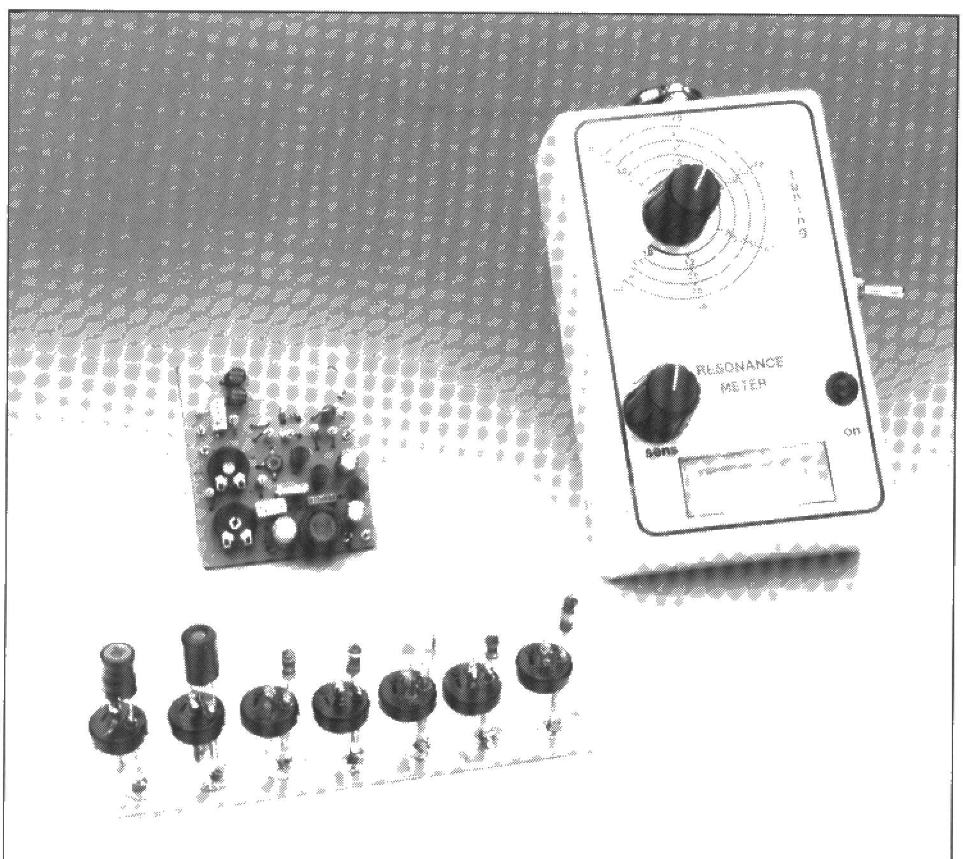
**A series of projects for the not-so-experienced constructor. Although each article will describe in detail the operation, use, construction and, where relevant, the underlying theory of the project, constructors will, none the less, require an elementary knowledge of electronic engineering. Each project in the series will be based on inexpensive and commonly available parts.**

## 5. Resonance meter

**J. Bareford**

**The present intermediate project follows this month's theme by venturing out into the fascinating world of radio techniques. The test instrument discussed is a must for anyone working with RF signals, but with a limited budget. It enables the resonance frequency of tuned circuits to be measured within the range of 100 kHz to about 50 MHz, and can also be used as a capacitance meter, RF test generator and RF signal probe.**

Traditionally, the name of the instrument of the type to be described has evolved from *grid dipper* to *gate dipper* or simply *dipper*. The first name, grid dipper, was used in the valve era and long after. When thermionic valves disappeared from consumer electronic equipment, the instrument was built from semiconductors and baptized 'gate dipper' because the gate of a field effect transistor (FET) is electrically very similar to the first grid of a valve. The instrument is basically an RF signal source with adjustable output frequency, coupled to a circuit that measures and indicates the amplitude of the output signal — see Fig. 1. Because the terms 'gate' and 'grid' have been formed historically, but have really nothing to do with the basic function of the instrument, these misnomers are omitted here to be replaced by the more universal term 'resonance meter'.




### Tuned circuits and resonance

Many constructors shy away from projects that contain home-made inductors, because these, they feel, remain something of a mystery owing to their lack of experience or suitable test and measuring equipment. And yet, many a radio amateur will confidently inform these constructors that there is nothing mysterious about winding coils. In fact, dimensioning them and peaking the resultant tuned circuit at the right frequency is sheer pleasure, provided, he will tell you, that a resonance meter is available. Without this simple instrument even experienced RF engineers are often at a loss in getting radio equipment to work correctly.

Any tuned circuit absorbs energy from another that is placed near it, and resonates at the same frequency. The RF energy is supplied by the resonance meter and an inductor that forms part of an oscillator. When this inductor is held near the coil under test, the oscillator output amplitude drops if the two tuned circuits resonate at the same frequency. When the 'dip' is indicated by the signal level meter on the resonance meter, the resonance frequency of the tuned circuit under test can be read from the tuning dial. Mind you:

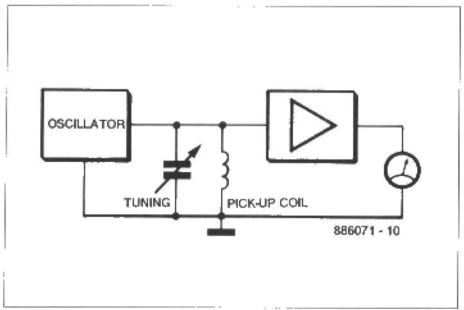



**Fig. 1. Block diagram of the resonance meter.**

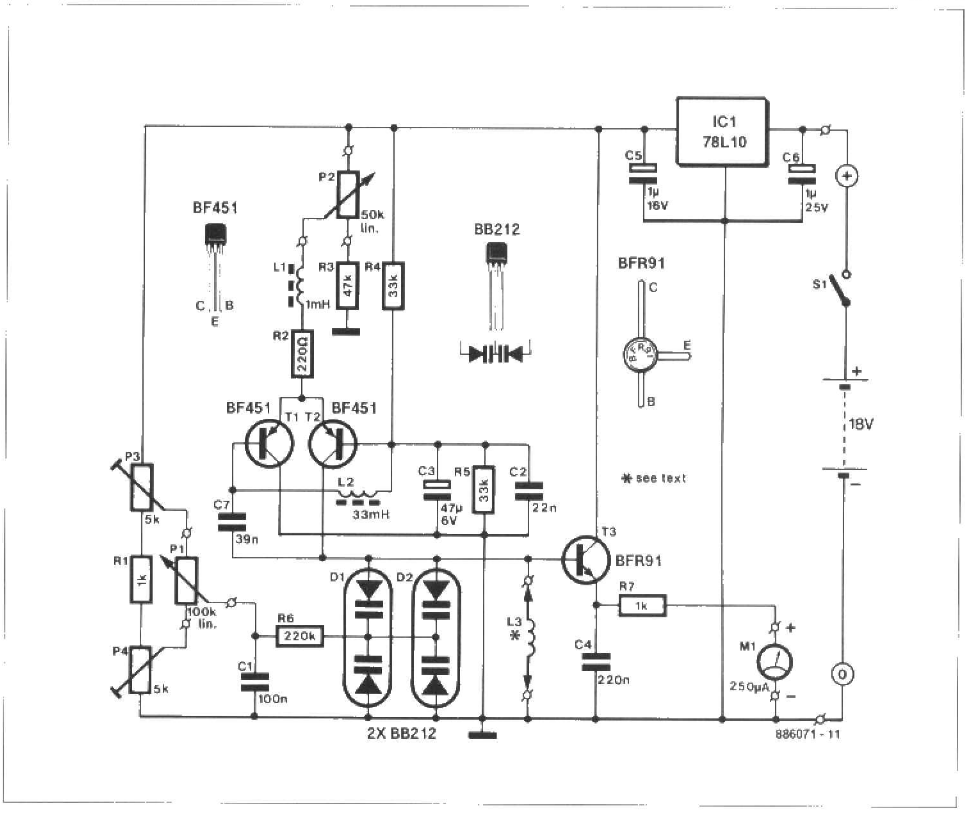



**Fig. 2. Circuit diagram of the budget resonance meter for 0.1–50 MHz in eight ranges.**

the resonance frequency can be determined while the equipment of which the tuned circuit forms part is not powered. The coupling between the resonance meter and the tuned circuit under test is entirely inductive: all that is required is to hold the pick-up coil on the meter close to the tuned circuit under test. Tune the resonance meter, and the signal level meter on it will tell you the resonance frequency of the *L-C* network under test.

## Resonance meter as an RF signal source...

Since the resonance meter contains an oscillator capable of covering a fairly large frequency range, it may double as an RF signal generator. To align a receiver, for instance, the resonance meter is simply set to the required frequency and placed close to the aerial input. If it is too strong for a precise adjustment, the test signal can be attenuated by placing the resonance meter further away.

## ... as a frequency meter or RF probe...

The resonance meter is designed such that it can easily be used as a coarse frequency meter and signal strength meter (RF probe). These functions are achieved by switching off the internal oscillator, but leaving the pick-up coil and the signal rectifier plus level indicator in function. Energy picked up from a resonating inductor in equipment to be aligned thus causes the meter to deflect when the tuning dial is set to the correct frequency. The meter indication is a measure of the signal strength, while the tuning dial shows the measured frequency. These combined functions are particularly useful for aligning receivers and transmitters in which several frequencies are mixed. The probe function of the resonance meter is then ideal for, say, aligning the filter that follows the mixer stage, so that only the wanted frequency is passed.

| $f$ (MHz) | $L$ (H) |
|---|---|
| 0.1–0.2 | 10m |
| 0.2–0.45 | 2m2 |
| 0.45–1.0 | 470µ |
| 1.0–2.0 | 100µ |
| 2.0–4.5 | 22µ |
| 4.5–10 | 4µ7 |
| 10–20 | 1µ |
| 15–40 | 0µ22 |

**Table 1. The values of $L_3$.**

## ...and a C or L meter

Capacitance (*C*) and inductance (*L*) measurements are the last additional functions of the resonance meter.

The value of a capacitor can be determined with the aid of a parallel inductor with known inductance, *L*, and, of course, a resonance meter. The capacitance, *C*, is simple to calculate from the resonance frequency, $f_0$, of the parallel tuned circuit:

$$f_0 = \frac{1}{2\pi\sqrt{LC}}$$

Since the self-inductance is known, and the resonance frequency can be measured, the equation can be rewritten as

$$C = \frac{1}{40 f_0^2 L}$$

Similarly, inductance can be calculated with the aid of a reference capacitor:

$$L = \frac{1}{40 f_0^2 C}$$

## Three transistors

The circuit diagram of the resonance meter is given in Fig. 2. All functions discussed above are realized by three transistors and a handful of passive components. Although perhaps a little difficult to deduce from the circuit diagram, $T_1$ and $T_2$ form an oscillator. The frequency of oscillation is determined by $L_3$ and varicaps $D_1$ and $D_2$. The two diodes are connected in parallel to achieve the required capacitance range that can be adjusted with $P_1$.

A total of eight plug-in inductors is required to cover the frequency range from 0.1 to 50 MHz.

Preset $P_2$ allows the collector current in both transistors to be adjusted, giving control over the amount of RF energy generated by the oscillator. Transistors $T_1$ and $T_2$ form a *differential amplifier* in which $C_7$ provides the feedback between the collector of $T_2$ and the base of $T_1$.

The measuring amplifier is formed by $T_3$. This transistor is operated in class C, so that it does not conduct until the voltage on $L_3$ is about 0.6 V higher than the emitter voltage. This means that $T_3$ forms a basic rectifier because it conducts only during a part of the positive half-wave of the oscillator signal. This pulsating signal is converted into a clean direct voltage by $C_4$. Regulator $IC_1$ prevents fluctuations of the supply voltage degrading the stability of the oscillator. Should the supply voltage be unstable, the voltage at $P_1$, and with it the varicap voltage, is unstable also. The varicap voltage, by the way, is not only dependent on the setting of $P_1$. Presets $P_3$ and $P_4$ are included to give $P_1$ the correct range, which is a must for the calibration of the scale on the front-panel designed

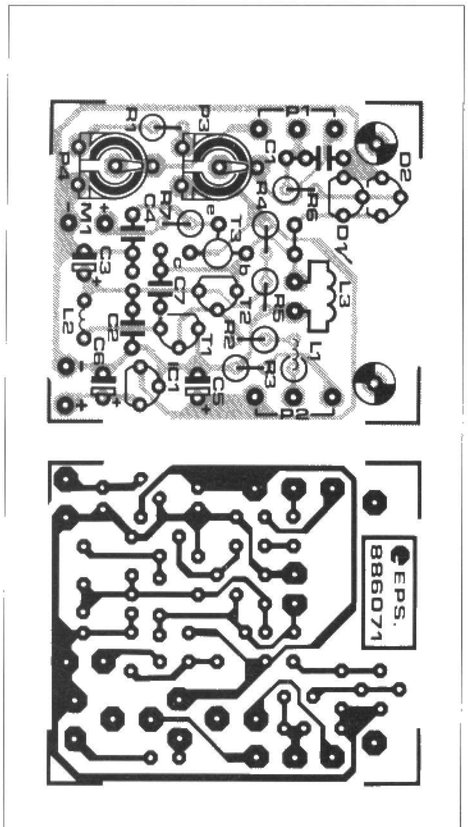
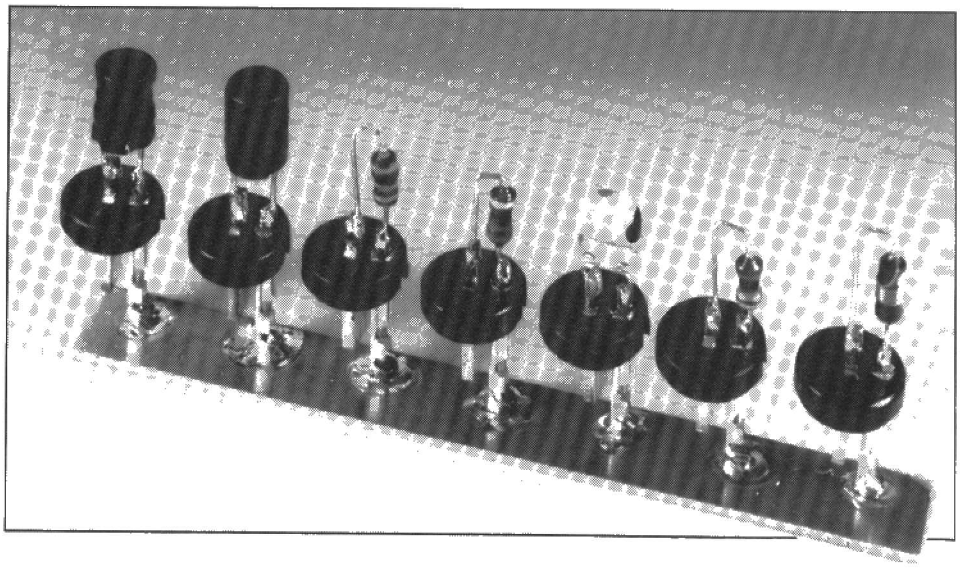


Fig. 3. Printed-circuit board for the resonance meter.



for the resonance meter. The ranges of the resonance meter can only be calibrated if the standard choke values listed in Table 1 are used.

The circuit diagram shows that the supply voltage for the resonance meter is 18 V, obtained from two series-connected 9 V batteries. A mains adaptor with 12 VDC output is, however, also suitable if the resonance meter is fitted with a Type 78L10 voltage regulator.

## Construction

Resonance meters for frequencies up into the VHF range are not the easiest of construction projects. The main problem of many home-made as well as ready-made meters is that these produce 'false' dips when the pick-up coil is not held near a tuned circuit. According to Murphy's law, these false dips will typically occur in the most frequently used ranges.

**Parts list**

**Resistors (±5%):**
$R_1$;$R_7$ = 1k0
$R_2$ = 220Ω
$R_3$ = 47k
$R_4$;$R_5$ = 33k
$R_6$ = 220 k
$P_1$ = 100k linear potentiometer
$P_2$ = 50k linear potentiometer
$P_3$;$P_4$ = 5k preset H

**Capacitors:**
$C_1$ = 100n
$C_2$ = 22n
$C_3$ = 47µ; 6 V; radial
$C_4$ = 220n
$C_5$ = 1µ0; 16 V
$C_6$ = 1µ0; 25 V
$C_7$ = 39n

**Semiconductors:**
$D_1$;$D_2$ = BB212
$T_1$;$T_2$ = BF451
$T_3$ = BFR91
$IC_1$ = 78L10

**Miscellaneous:**
PCB Type 886071 (see Readers Services page).
$L_1$ = 1mH0 radial choke (Toko).
$L_2$ = 33mH radial choke (Toko).
$L_3$ = see Table 1
$S_1$ = miniature SPST switch.
$M_1$ = 250 µA moving-coil VU meter.
1 off DIN-type 2-way loudspeaker socket.
8 off DIN-type 2-way loudspeaker plugs.
ABS enclosure.

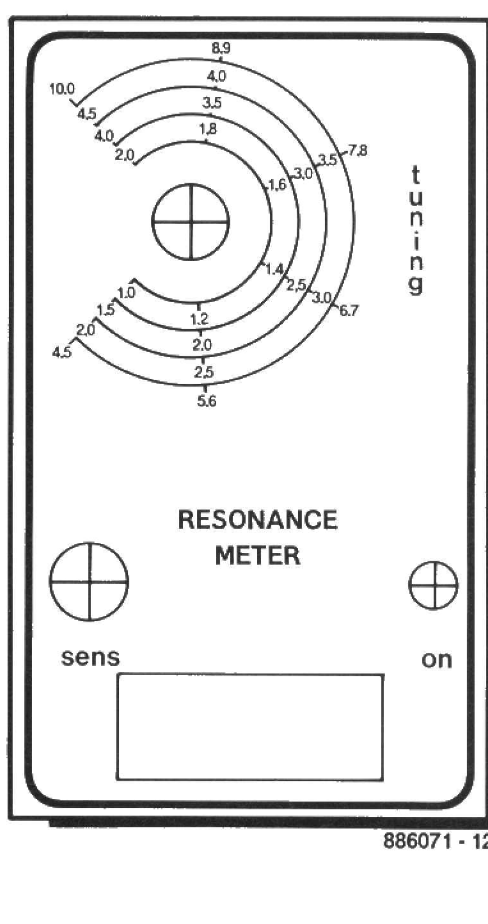



Fig. 4. True-size front-panel layout.

The printed-circuit board for the present resonance meter has been designed to minimize the risk of false dips. Figure 3 shows the component mounting plan and the track layout of the board, which is available ready-made.

Mounting the parts on the board is fairly straightforward. The only point to pay special attention to is that all component wires must be kept as short as possible.

The populated board is mounted in an ABS enclosure. This is not standard practice in view of RF screening, but avoids the risk of a metal enclosure affecting the operation of the oscillator. As a result, false dips would occur, and the calibration would have to be changed.

All wires in the enclosure, and particularly those between the board an the inductor socket should have the absolute minimum length. The front-panel design shown in Fig. 4 is copied and secured on to the enclosure.

The pick-up coils are made from DIN-type 2-way loudspeaker plugs and ready-made chokes as shown in the photograph of Fig. 5. The chokes for the two lowest ranges must be types with plastic sleeving, **not** types with ferrite encapsulation. The other chokes are miniature axial types.

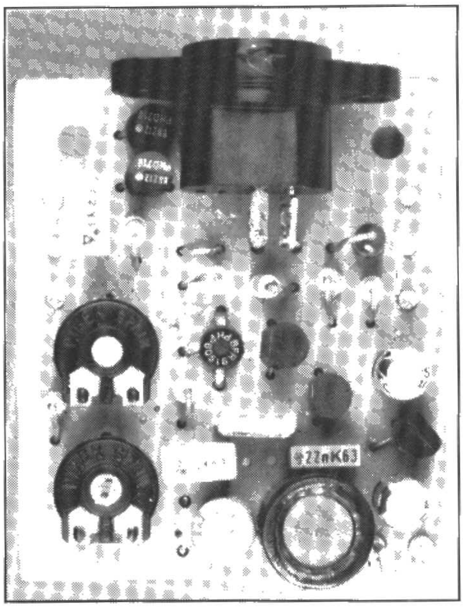
## Calibration

The resonance meter can not be calibrated before it has been fitted into a suitable enclosure. Either a frequency meter or a short-wave receiver must be used for the adjustment procedure.

If a frequency meter is available, the procedure is started by winding 10 turns of enamelled copper wire on to a lead pencil. Remove the pencil, and connect the inductor to the input of the frequency meter. Plug one of the lower-range coils into the resonance meter, switch on the instrument, and adjust $P_2$ for full-scale deflection of the signal level meter, $M_1$. The frequency meter will display a frequency if the pick-up coil on the resonance meter is held near that on the frequency meter. Check whether the displayed frequency rises if $P_1$ is turned anti-clockwise. If not, swap the outer wires on the potentiometer. Set the tuning to the highest frequency in the range, and adjust $P_3$ until the frequency meter displays the scale frequency. Turn $P_1$ to the lowest frequency, and adjust $P_4$ similarly. Once again check the upper frequency and correct the setting of $P_3$ if necessary.

A short-wave receiver is also suitable for calibrating the resonance meter, but has the disadvantage of requiring to be re-tuned for every adjustment.

Since every choke has its particular tolerance, it is necessary to check for scale deviations in every range of the resonance meter. If the deviation in a particular range is unacceptable, try using another choke from another batch but with the same value indication. Choke tolerance is typically ±20%.




## Practical use

The resonance meter is a test instrument that becomes easy to operate only gradually through regular practical use. Prior to any measurement, the frequency range must be determined, and the appropriate pick-up coil selected. In some cases, you will need to change coils if the resonance frequency is close to the end of the range. Switch on the instrument, and adjust the sensitivity control, $P_2$, for f.s.d. (full-scale deflection) of the signal level meter. Line up the pick-up coil with the inductor in the equipment (Fig. 6), and tune carefully until the pointer of the level meter moves to the left. The frequency range is probably fairly large at this stage. To achieve a more accurate dip, move the pick-up coil away from the inductor while still ensuring that they point in the same direction. Now retune the resonance meter until a sharp dip is found.

If the resonance frequency of a tuned circuit is not known, it is wise to start examining it in the lowest range of the resonance meter, increasing the range until a sharp dip is found. This procedure avoids harmonics being mistaken for the natural frequency.

The resonance meter need not be very accurate since its main application is the coarse dimensioning and adjustment of inductors, or capacitors that form part of an *L-C* tuned circuit — precise adjustment is invariably done along the lines of the setting-up procedure with the equipment turned on. Also, it is useful to note that the resonance frequency of an *L-C* circuit is generally lowered when it is installed in the circuit, which introduces additional capacitance.

Not all tuned circuits can be tested with the aid of the resonance meter. Inductors wound on a toroid core, or enclosed by a metal cover, absorb very little externally applied energy, and do not produce a dip unless a small external series inductor of one or two turns is added temporarily. This lowers the resonance frequency to some extent, but allows a useful estimate to be made. Some in-circuit *L-C* networks will not dip either. Examples are the heavily damped tuned circuits in the emitter line of a grounded-base transistor circuit. To measure the resonance frequency, either the transistor or the tuned circuit must be removed.

Finally, the resonance frequency of series *L-C* tuned circuits can not be measured unless a capacitor is included in the circuit that provides a path from the inductor back to the series capacitor. This, in fact, creates a parallel tuned circuit.

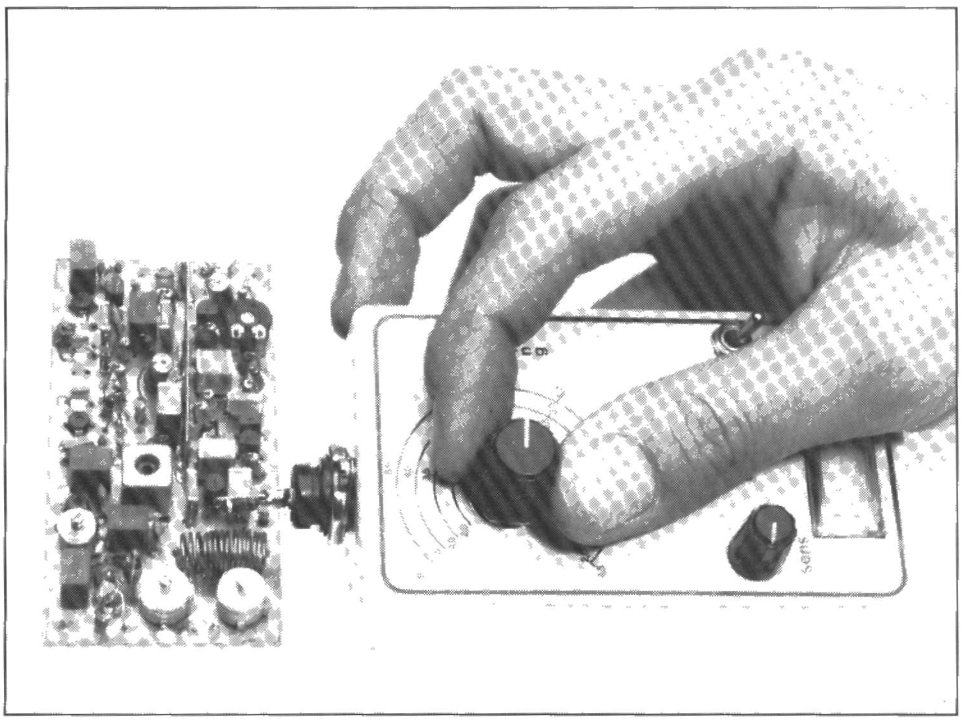


**Fig. 6. Using the resonance meter to 'dip' an *L-C* tuned circuit.**

# CENTRONICS MONITOR

**A. Rigby**

**The Centronics interface standard is often said to be without problems in practical use. When a malfunction occurs, however, many more connections have to be checked than, for instance, on an RS-232 link. The monitor described here alleviates the plight in debugging a Centronics connection. It is a handy tester that indicates the levels on all lines simultaneously, including those that carry pulses.**

Nearly all of today's personal computers are equipped with a Centronics port for connecting a printer. The general acceptance of the Centronics interface standard has been helped by the availability of ready-made cables of various lengths, and the fact that the majority of printer manufacturers have ensured compliance with the pinning of the 'blue-ribbon' input connector on their products.

Sometimes, however, the installation of a new printer or cable gives rise to awkward problems that take a lot of precious time to analyse and resolve. In these cases, the present in-line indicator provides almost instant fault analysis because it shows the logic state of the databits and a number of handshaking and control signals.

## Data and handshaking

To ensure that the computer-to-printer link works as required, a number of signals must be present, while the use of others depends on the equipment used at either side of the Centronics cable. At the computer side, datalines D0–D7 must be connected, as well as handshake lines $\overline{\text{STROBE}}$, BUSY and/or $\overline{\text{ACK}}$ (*acknowledge*). Especially the last two signals are prone to cause trouble if the relevant pins are correctly marked (according to the printer manual), but not used electrically.

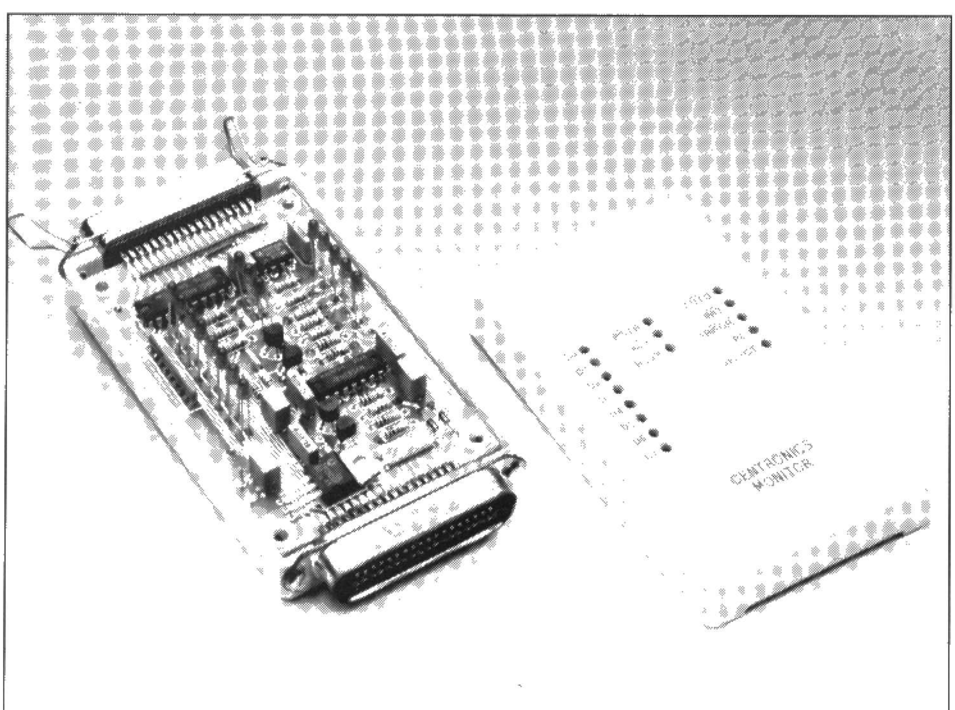
When the Centronics connection is fully functional, the computer puts the bit pattern to be sent to the printer on to the eight datalines, and actuates the $\overline{\text{STROBE}}$ line by pulling it low. This enables the printer to recognize that the databyte representing the printable character is stable and therefore valid. Reception of the byte is signalled to the computer by a low-to-high change on the BUSY line. BUSY remains high until the printer is ready. Depending on the type of printer, the received databyte is instantly printed, or stored in an internal buffer memory. In both cases, however, the *processing* (which is not necessarily the same as *printing*) of a character is signalled to the computer by means of a high-to-low transition on the $\overline{\text{ACK}}$ line.

The processing of received characters differs from printer to printer. Older models print each character immediately after it has been received, halting the computer during the printing operation. Printers of a later generation typically feature a small buffer that allows a line of printable characters to be stored. The characters in this buffer are not printed until a carriage return is received. Many modern matrix printers have buffers capable of storing many kilobytes of text, and handle printing, data spooling and communication with the computer simultaneously. Some top-range models house more data processing chips than the average PC compatible.

Apart form the data and handshaking lines, the Centronics standard specifies a number of other, auxiliary, functions:

| | | |
|---|---|---|
| PE | Paper Empty | Goes high when the printer is out of paper. |
| SEL | Select | Indicates that the printer is on line and ready to receive data. |
| $\overline{\text{AUTO}}$ | Auto Feed | Automatic line feed after a carriage return. |
| $\overline{\text{INIT}}$ | Initialize | Resets the printer |
| ERROR | | Indicates internal failure. |

The last four lines must be given a fixed level, even if they are not used in the actual connection between the computer and the printer. In other words: the minimum requirement is that non-connected active-low and active-high lines be fitted with a pull-up and pull-down resistor respectively.



## The monitor

The Centronics monitor indicates the current logic level on all lines by means of light-emitting diodes (LEDs). The databus lines are connected to a Type 74HCT540 buffer that supplies sufficient output current to connect the LEDs direct to ground. A logic high level causes the LED associated with a particular line to light. Each of the five status lines is connected direct to the associated LED. This can be done with impunity because the signal levels are

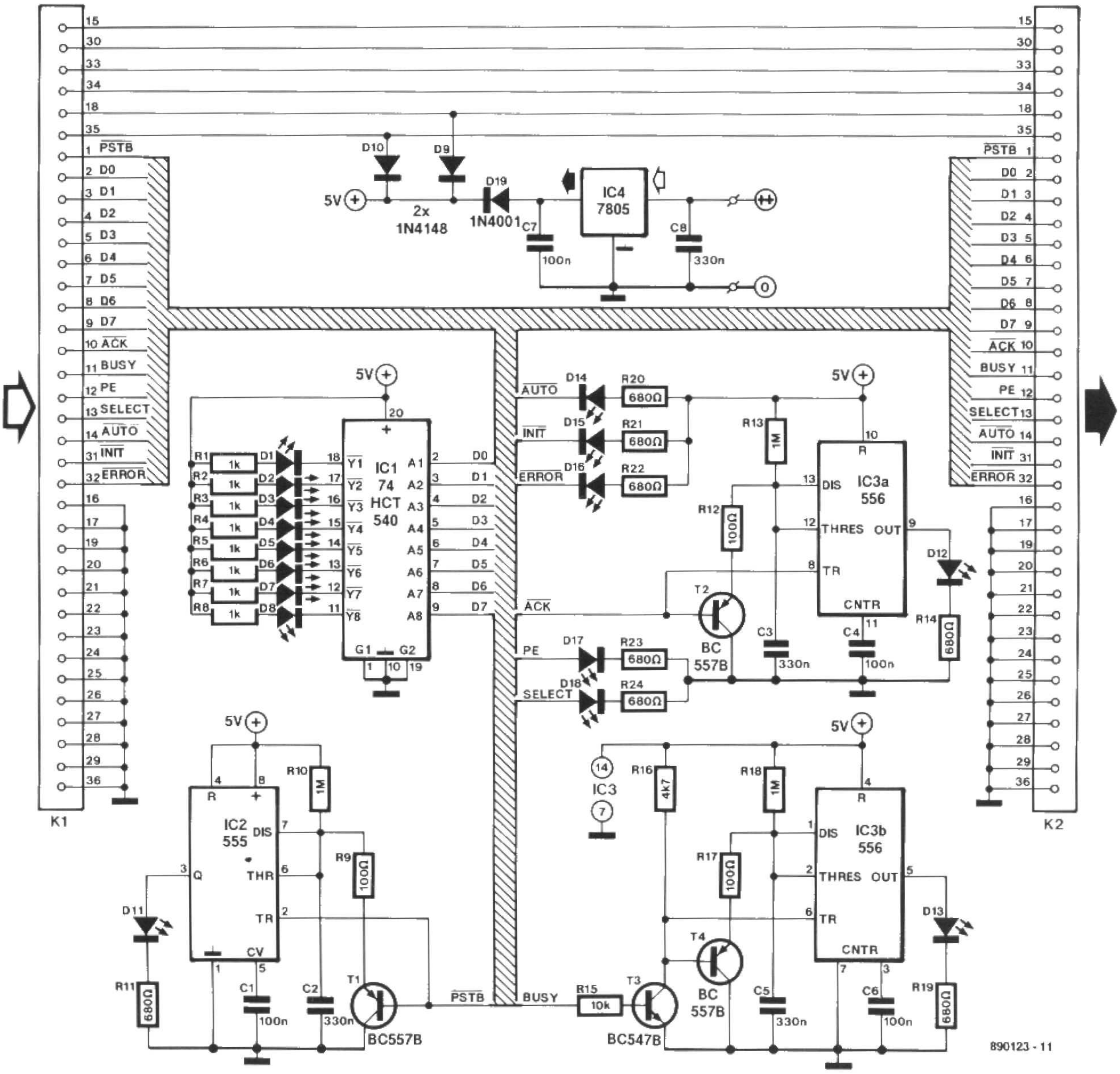



**Fig. 1. Circuit diagram of the Centronics monitor.**

fairly steady.

Signal lines $\overline{\text{STROBE}}$, BUSY and $\overline{\text{ACK}}$ require a different configuration because they carry pulses rather than steady levels. Three monostables in the form of Type 555 timer chips are therefore used to drive the relevant LEDs. Inverter $T_3$-$R_{15}$-$R_{16}$ ensures that the BUSY LED lights when the associated line is actuated (i.e., logic high). Such an inverter is not required for the $\overline{\text{ACK}}$ and $\overline{\text{STROBE}}$ lines, which are active-low. The associated 555s are housed in dual timer $IC_3$, a 556.

All other signals that may be available, but are not strictly required for correct operation, are simply passed between the relevant pins of the input and output socket of the monitor. Many Centronics cables do not have separate ground wires, but use commoned connector pins at both ends. These pins are often connected by a single wire.

Sixteen LEDs enable the user of the monitor to locate the possible source of trouble at a glance: 8 data LEDs, 3 for the handshaking lines, and 5 for the status lines.

## Power supply

An external supply will not be required in most cases because virtually all modern printers supply +5 V at pin 18, 35 or both. Diodes $D_9$ and $D_{10}$ ensure compatibility of the monitor with these printers, and also allow the unit to be powered from an external 10 V/50 mA power supply. Regulator $IC_4$ then provides the 5 V supply voltage for the ICs and LEDs on the board.

## Connections

Connector K1 is a 36-way Centronics socket with straight solder pins. Push the socket on to the PCB edge, while ensuring that the pins align with the copper islands. Soldering is then straightforward. Connector K2 is removed from a standard Centronics cable plug, and secured as K1. Two types of connector exist: versions with screws and versions with clamps for the screening hood. The screw type is the better for the present application.

| Pin | Signal | Source |
|---|---|---|
| 1 | STROBE | Computer |
| 2 | Data 0 | Computer |
| 3 | Data 1 | Computer |
| 4 | Data 2 | Computer |
| 5 | Data 3 | Computer |
| 6 | Data 4 | Computer |
| 7 | Data 5 | Computer |
| 8 | Data 6 | Computer |
| 9 | Data 7 | Computer |
| 10 | ACK | Printer |
| 11 | BUSY | Printer |
| 12 | PAPER EMPTY | Printer |
| 13 | SELECT | Printer |
| 14 | AUTO FEED XT | Computer |
| 15 | n.c. | |
| 16 | ground | |
| 17 | chassis | |
| 18 | +5 V | Printer |
| 19 | ground | |
| 20 | ground | |
| 21 | ground | |
| 22 | ground | |
| 23 | ground | |
| 24 | ground | |
| 25 | ground | |
| 26 | ground | |
| 27 | ground | |
| 28 | ground | |
| 29 | ground | |
| 30 | ground | |
| 31 | INIT | Computer |
| 32 | ERROR | Printer |
| 33 | n.c. | |
| 34 | n.c. | |
| 35 | +5 V | Printer |
| 36 | n.c. | |

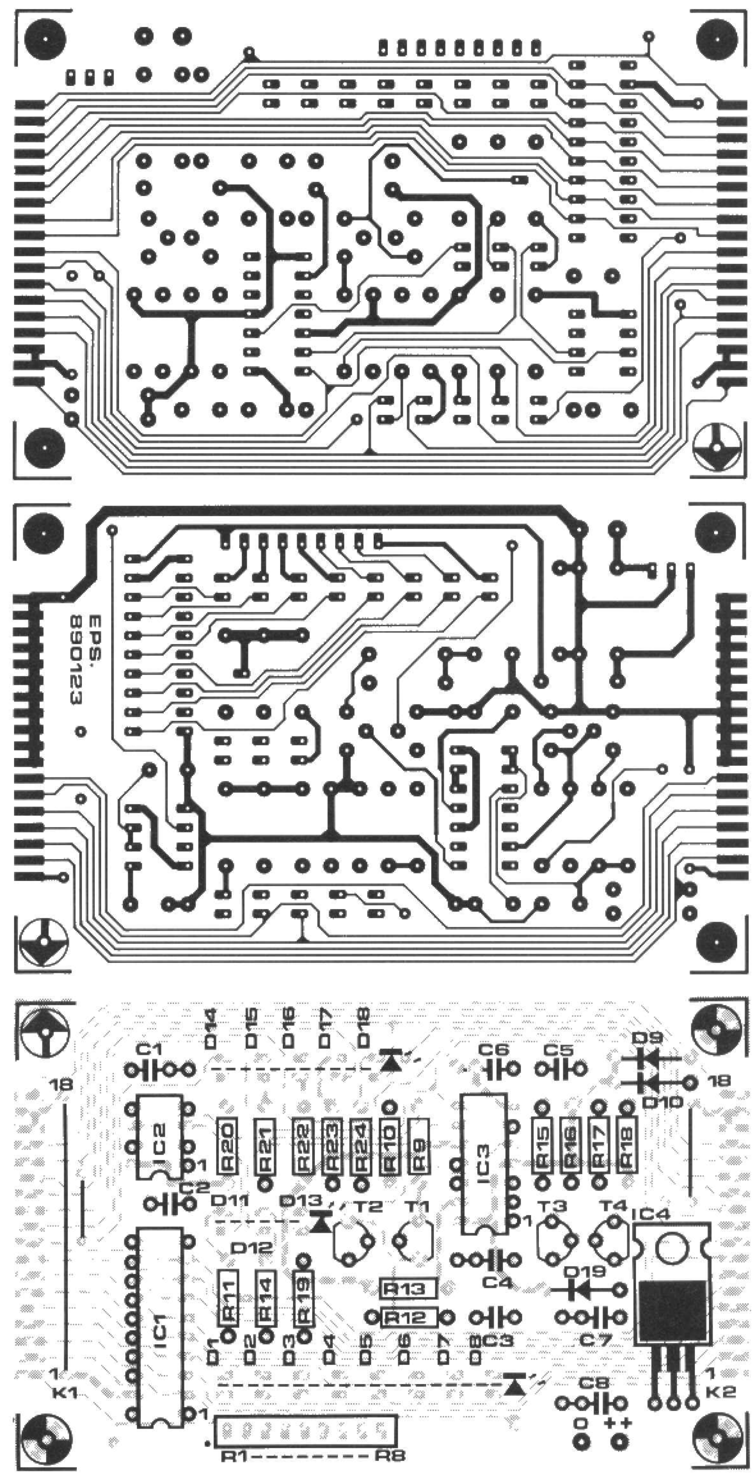



Fig. 2. Track layout and component mounting plan. A number of parts must be soldered at both sides of the printed-circuit board.

**Parts list**

**Resistors** (±5%):
R1–R8 = 1k0 SIL resistor array
R11;R14;R19–R24 = 680Ω
R9;R12;R17 = 100Ω
R10;R13;R18 = 1M0
R15 = 10k
R16 = 4k7

**Capacitors:**
C1;C4;C6;C7 = 100n
C2;C3;C5;C8 = 330n

**Semiconductors:**
D1–D8;D11–D18 = LED; 3 mm; red
D9;D10 = 1N4148
D19 = 1N4001
T1;T2;T4 = BC557B
T3 = BC547B
IC1 = 74HCT540
IC2 = 555
IC3 = 556
IC4 = 7805

**Miscellaneous:**
K1 = 36-way Centronics socket with straight solder pins.
K2 = 36-way Centronics plug.
Enclosure: e.g., OKW model A9407113.
PCB Type 890123 (see Readers Services page).

# ASIC MICROCONTROLLERS

by Simon Young*

**This article discusses the evolution of the MCS-51 architecture and how to use ASIC technology to extend the set of generic features contained in the family members.**

Intel offers the MCS-51 architecture to customers in a number of ways.

The first is via standard products, such as the 80C51BH and 8052. These devices are designed for the general microcontroller market, where the internal hardware resources can be closely matched to the system requirements.

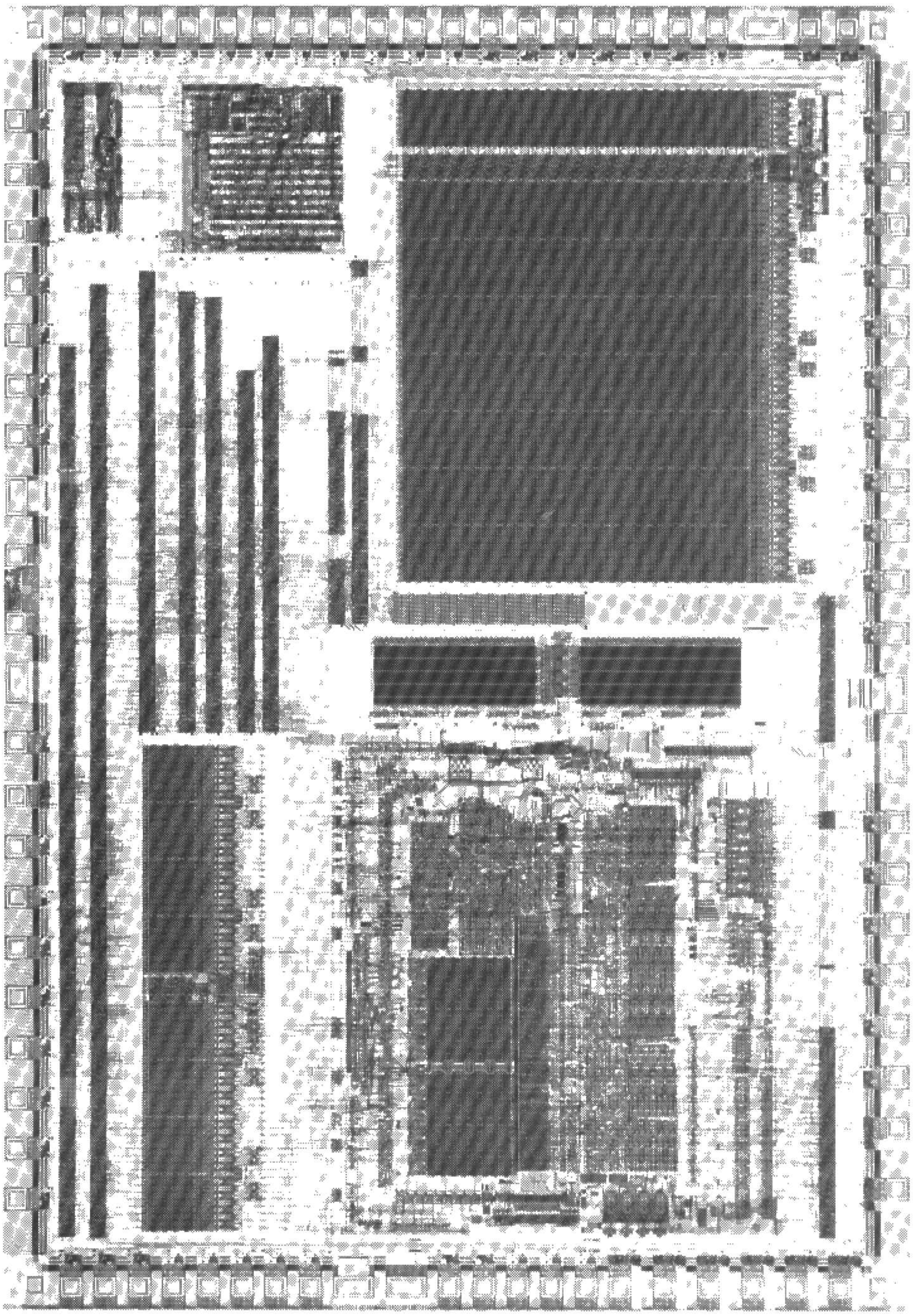
The second is via Application Specific Standard Products (ASSPs) developed for a vertical market sharing a common set of additional features. An example is the 80C51FA, which augments the MCS-51 core features with a programmable counter array, an enhanced serial port for multiprocessor communications and an up/down timer/counter.

The third way to gain access to the MCS-51 architecture is via ASIC, and this is the subject of this article. Intel offers to customers the same capability it uses in house to develop ASSPs.

## MCS-51 Microcontrollers

The MCS-51 family of microcontrollers was designed to meet the needs of embedded control applications. The architecture and instruction set were optimized for the movement of data between internal memory and internal peripherals.

Figure 1a shows the MCS-51 architecture. The Special Function Register (SFR) bus connects the internal resources (such as port latches, timers and peripheral control registers) with the CPU. The128 bytes of on-chip RAM (between 00 hex and 7F hex) can be addressed both with direct (MOV data addr) and indirect (MOV @Ri) addressing modes. Some devices, e.g., the 8052, provide an additional 128 bytes of on-chip RAM for temporary data storage between 80 hex and FF hex (dotted in Fig. 1b). This may be addressed only indirectly – forming a useful area for the stack.



Typical 80C51 core-based design.

The SFR space appears to the CPU as 128 bytes of memory located between 80 hex and FF hex. This area of memory is accessed only by direct addressing modes, in order to distinguish it from the additional data RAM discussed above. Of the 128 locations, 21 are used in the 80C51BH standard product (26 on the 8052).

The 64 Kbytes of external data memory space are accessed with the MOVX instruction.

Another powerful feature of the MCS-51 architecture is the ability to address individual bits within certain SFR and internal RAM locations. All MCS-51 devices contain a complete Boolean (single-bit) processor. The MCS-51 instruction set supports the Boolean processor with instructions to move, set, clear, complement, OR, AND, and conditional branch on bit. This 'bit addressability' allows individual bits to be tested and modified without the need of complex masking operations, with consequent significant improvements in speed.

*Simon Young is with Intel Corporation (UK) Ltd at Swindon

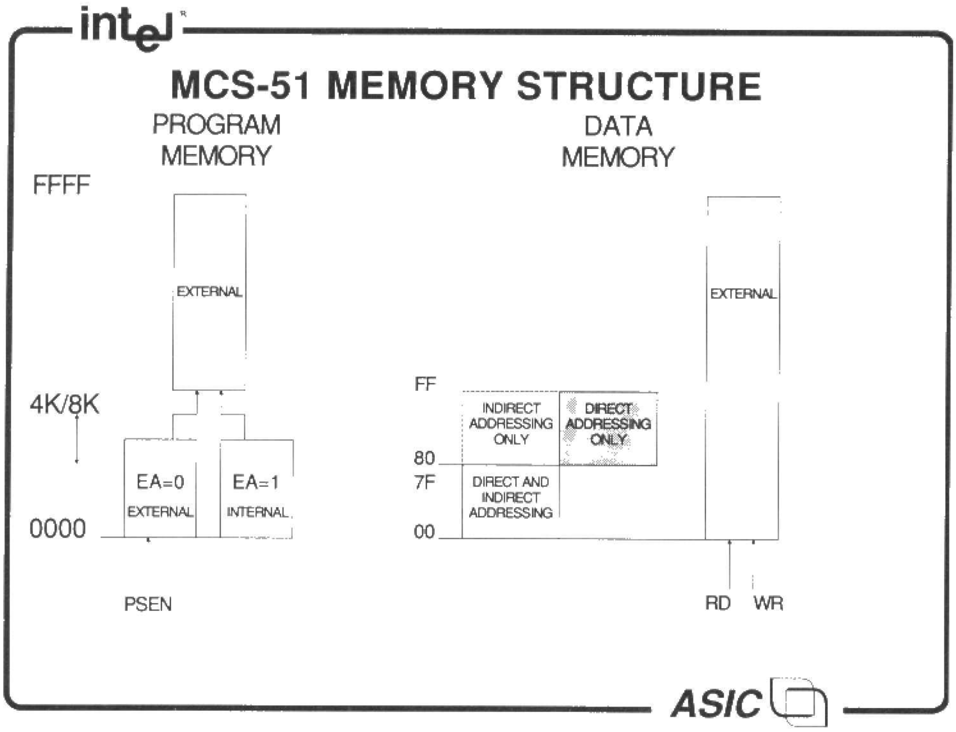



Fig. 1a

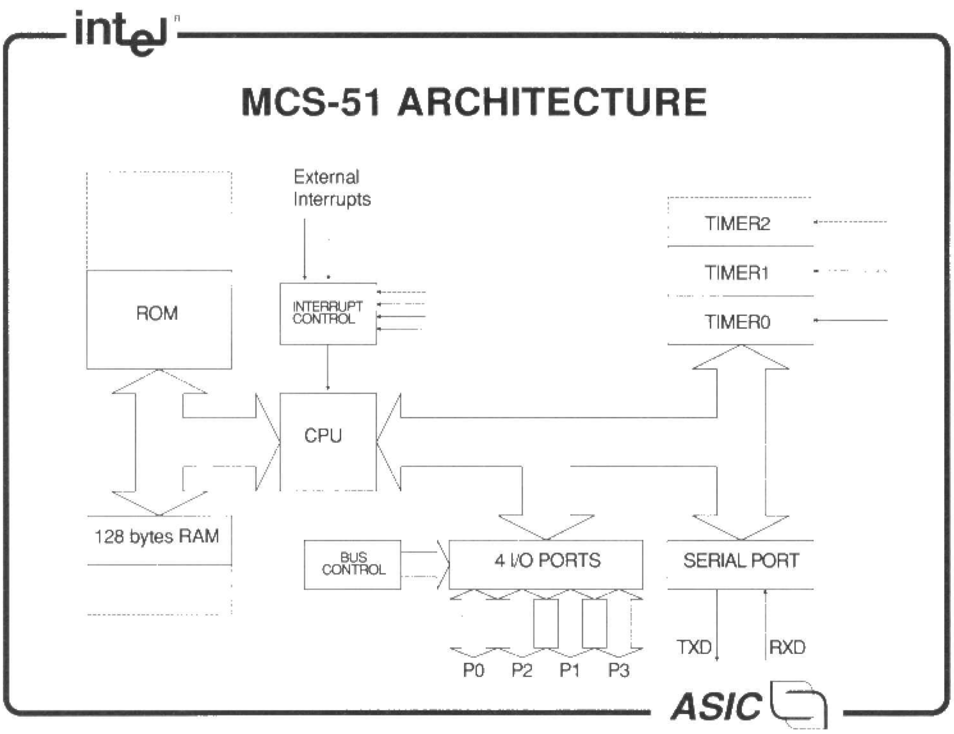



Fig. 1b

At the time the first members of the MCS-51 family were introduced, there was no economic packaging for high pin-count devices. The parts were packaged in 40-pin DIL packages: only recently have PLCC packages been used. To provide access to the internal hardware resources of the device, several functions, including port input and output signals, external multiplexed address/data bus, serial port I/O, external interrupt signals and timer/counter input signals had to be multiplexed. Clearly, functions multiplexed on the same pin may not be used concurrently.

As systems designers developed increasingly complex embedded control applications, the 8051 required additional memory, peripherals and/or I/O ports. These had to be added externally as memory or memory-mapped peripherals, reducing the parallel I/O available on the 8051. Fully expanded in this way, only a single 8-bit I/O port is available. While the on-chip features and price-performance ratio of the 8051 make it still an attractive proposition when compared with other solutions, the end result is different from what the 8051 was designed to be: a single-chip, stand-alone microcontroller.

## UC51: Intel's original microcontroller core

In 1985, Intel introduced the UC51, which was developed from the 1.5 μm CHMOS III 80C51BH standard product. The UC51 allows designers to integrate the microcontroller core, memory, memory-mapped peripherals and cells from the 1.5 μm standard cell library on to a single chip.

In transforming the 80C51BH into the UC51 core cell, the I/O pads and pin multiplexers were removed. The internal peripherals, multiplexed address-data bus, control signals and input and output ports all have dedicated signals.

With the ability to choose different amounts of program ROM (zero, 4 K, 8 K or 16 K bytes) and data RAM (up to 1 K bytes) with no loss in functionality, the UC51 has been a very successful part of Intel's ASIC offering.

In summary, the UC51 provides systems designers the capability to integrate a 'fixed' core and memory-mapped peripherals, complete with user-defined logic, on to a single ASIC device. The ASIC resembles an integrated version of the discrete solution, with increased flexibility because of the demultiplexed I/O. However, it is not possible to apply the full power of the architecture and instruction set to memory-mapped peripherals.

## UCS51: Intel's next generation microcontroller core

Intel have recently introduced the UCS51 family of microcontroller and peripheral cells into the 1.5 μm CHMOS III standard cell library. The UCS51 permits systems designers to connect any of the available peripheral cells or user-defined logic directly into the SFR space. The UCS51 cores then access the control registers within these peripherals in exactly the same way as any internal SFR register.

There are great benefits to be gained from directly connecting peripherals to the SFR bus:

- instructions operating on peripheral registers in the SFR space are more code-efficient then accessing memory-mapped registers indirectly (with MOVX), so that less program memory space is required;
- register-direct-instructions (ADD, ADDC, SUBB, INC, DEC, ANL, ORL, XRL, MOV, PUSH, POP, XCH, CJNE and DJNZ) execute more quickly, giving improved system throughput;
- certain bytes in the SFR space (located at x0 hex and x8 hex) are bit addressable: mapping peripherals into these locations permits the bit-banging capabilities of the Boolean processor to be applied to these registers;

• interface logic between the UCS51 core and UCS51 peripherals is eliminated: there is no need of an address latch, address decoder or tri-state bus driver.

The basic UCS51 core cell resembles a UC51. PORT1 has been removed to provide access to the SFR bus, although it may be replaced easily as described later. Interfaces have been added for connecting ROM modules (either none or one of 4 K, 8 K, or 16 K bytes), a RAM module (same RAM as 8052) and an interrupt expansion unit. A functional cell diagram is shown in Fig. 2.

## Additional interrupts enhance real-time performance

Unexpanded UCS51 cores have five interrupt signals available, as have the UC51 and 80C51BH. Users may configure the internal peripheral interrupts for use as general purpose interrupt signals, with no change in priority levels and vector locations. It is also possible, by the use of the Interrupt Expansion Unit, to add a further five external interrupts, making a total of 10, with complete flexibility of interrupt source, peripherals, on-chip or off-chip logic.

## LSI peripherals for configuring unique microcontroller cells

The Bus Interface Unit is, perhaps, the most important UCS51 peripheral cell. Functionally, it is and 8-bit input, 8-bit output SFR bus interface. With it, designers may replace PORT1 and add further demultiplexed I/O ports as needed.

This cell is also used to interface between on-chip user-defined logic and the SFR bus. Thus, customer developed logic using cells from the 1.5 μm standard cell library may be mapped directly into the SFR space to gain the advantages discussed previously.

The 8-bit, 8-channel successive approximation ADC has a nominal conversion speed of 20 μs at a core frequency of 16 MHz. A conversion may be triggered by hardware or software, with an interrupt generated on completion.

Timer2 is a 16-bit timer/counter cell, enhanced over the Timer2 found on the 8052 standard product and some ASSPs.

The serial I/O cell is a full-duplex serial port, enhanced from the serial channel contained in the 80C51BH by the addition of a new mode: Mode 4. This mode provides 9-, 10-, 11- or 12-bit transfers with variable baud rate. In Mode 4, the UCS51SIO cell also generates parity for transmission and detects framing, overrun and parity errors on reception.

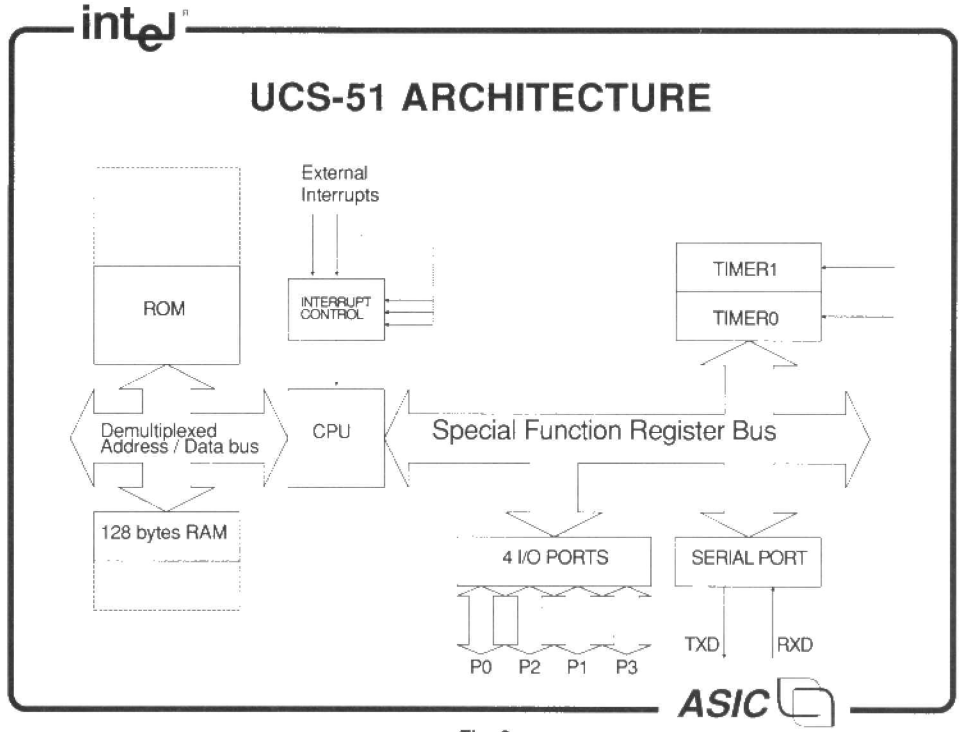



Fig. 2

The baud rate generator cell is used to generate clocks for the serial I/O peripheral or for user-defined logic. Operating from the 16 MHz system clock, the BRG generates rates from 50 Hz to 4 MHz with an accuracy better than 0.2%.

These five LSI peripheral cells and 16 distinct core configurations complete the UCS51 offering at the time of launch: more are in development. The 1.5 μm standard cell library includes SSI, MSI and I/O functions and may also be integrated on a UCS51-based ASIC.

## CAD tools aid development of microcontroller-based ASICS

The Design Entry Tool, DET, provides a high-level, menu-driven means of configuring the core hardware resources. The UCS51 core options are RAM, ROM or interrupts. Adding a peripheral requires two data to be entered: the peripheral type and the address of its control registers in the SFR space.

The DET outputs a symbol for this core. The designer simply adds the user-defined logic he requires, surrounds this with the I/O pads and the design capture is complete.

Once captured, the designs netlist is transmitted to MDVS – Intel's Mainframe Design Verification System based on VAX/ZyCad hardware. Full-timing gate-level simulation of the entire chip is possible with vectors written in TDPL – Intel's Test Pattern Development Language – and ExtASM51. ExtASM51 provides 8051 assembly source code and simulation stimulus, and synchronizes the execution of instructions with external stimulus applied to the ASIC.

The simulation output may be viewed as text on the host, or returned to the workstation for display and review in the graphics condition.

The ICE-UCS51 In Circuit Emulator allows the designer to develop and test code for a UCS51-based ASIC, and to emulate the completed ASIC (core, peripherals and user-defined logic) in the target system. The ICE is a PC-based emulator system, offering the same advanced features as Intel's other ICE systems.

The UCS51 core is tested with a slightly modified version of the 80C51BH standard product test program, guaranteeing functional and parametric equivalence to the standard part. The peripherals are tested in the same way.

The designer is responsible only for his user-defined logic, and provides TPDL and ExtASM51.

The result is standard producty quality and reliability: an AQL of 0.1% is guaranteed.

## Summary

Intel's family of UCS51 core and peripheral cells provides the systems designer with unprecedented flexibility in ASIC design. Not only is access provided to the basic core architecture of the 8051, but also to a specialized set of peripheral cells. The design tools guide the designer through design capture, simulation and test vector developments. All components of the UCS51 family were developed with one overriding aim: to provide guaranteed success of UCS51-based ASIC devices.

# APPLICATION NOTES

**The contents of this article are based on information obtained from manufacturers in the electronics industry, and do not imply practical experience by *Elektor Electronics* or its consultants.**

# HAYES-COMPATIBLE V22bis MODEM

**source: Silicon Systems, Inc.**

**The SSI73D2404 is a set of three chips from Silicon Systems Inc. that provides the data pump function required to construct a high-performance 2400bps full-duplex intelligent modem for use over the dial-up telephone network. The chip set includes operating modes compatible with CCITT V22bis, V22, V21, as well as Bell 212A and Bell 103 data communications standards.**

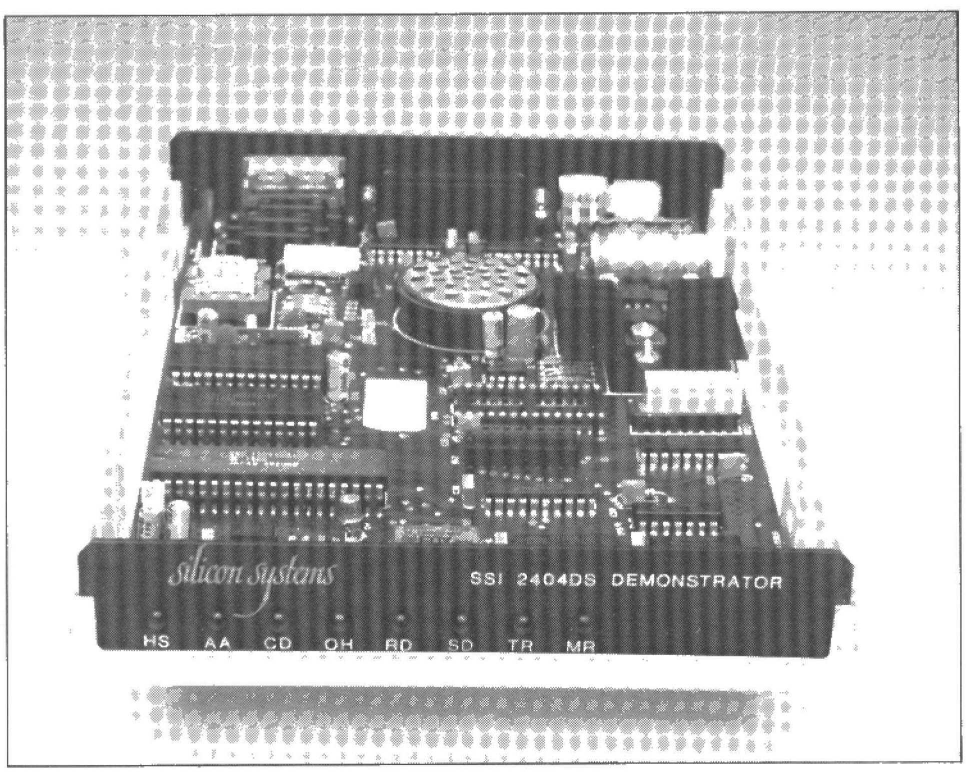
Using advanced CMOS processes that include analogue, digital signal processing and switched capacitor techniques, SSI's 73D2404 is claimed to offer excellent performance and a high level of functional integration in a compact three-chip set that includes one 40-pin and two 28-pin dual-in-line (DIL) packages.

The chip set is ideal for use in stand-alone as well as integral system modem products where full-duplex 2400 bit per second (bps) data communications over the two-wire public telephone network is desired.

## Operation

The SSI 73D2404 is a complete V22bis intelligent modem contained in three CMOS ICs. The device set forms the basis for a stand-alone modem with self-contained command interpreter, indicator LEDs, default switches and interface lines for an RS232 serial port. Both data and commands are passed over the serial port as customary for modems in the PC environment.

The SSI 73D2404 provides the QAM (quadrature amplitude modulation), PSK (phase shift keying) and FSK (frequency shift keying) modulator/demodulator functions, call progress and handshake tone monitors, test modes and a tone generator capable of producing DTMF (dual-tone multi-frequency), answer and CCITT guard tones. This device supports the V22bis, V22, V21 and Bell 212A/103 operating modes, both synchronous and asyn-




SSI MODEM CHIP SET

- Multi-mode V22bis, V22, V21 and Bell 212A/103 compatible device set for intelligent modem designs
- Full-duplex operation at 300, 1200, and 2400 baud with all operating modes both synchronous and asynchronous
- Includes high-level 'AT' command interpreter compatible with 2400 baud industry standard products
- Complete complement of 'AT' modem features
- Selectable automatic speed select, handshake and baudrate detect functions
- Supports external non-volatile memory to store user configurations
- Adaptive equalization for optimum performance over all lines
- Dynamic range from 0 to −45 dBm
- Call progress, carrier and answer tone detectors provide intelligent dialling functions
- DTMF and CCITT guard tone generators
- Test modes available: ALB, DL, RDL for complete test capability
- Space-efficient 28- and 40-pin DIPs
- All-CMOS technology for low-power consumption (<600 mW at 5 V)

chronous. The 73D2404 is designed to provide functions needed for an intelligent modem, and includes auto-dial and auto-answer, handshake with auto-fallback, and selectable pulse or DTMF dialling sequences to simplify these designs.

The chips set consist of three devices. The SSI73M214 is an analogue processor that performs the filtering, timing adjustment, level detection and modulation functions. The 73D215 is the receiver digital signal processor. The 73D216 is a command processor that provides supervisory control and command interpretation.

## QAM modulator/demodulator

The SSI73D2404 scrambles and encodes the 2400bps incoming data into quad bits represented by 16 possible signal points as specified by CCITT recommendation V22bis. The modulator transmits this encoded data using either a 1200 Hz (originate mode) or 2400 Hz (answer mode) carrier. The demodulator reverses this procedure and recovers a data clock from the incoming signal. Adaptive equalisation corrects for different line conditions by automatically changing filter parameters to compensate for line characteristics.

## PSK modulator/demodulator

In PSK mode, the 73D2404 modulates the 1200bps incoming data using a subset of the QAM signal points as specified by CCITT recommendation V22bis, V22 and Bell 212A. The PSK demodulator is similar to the QAM demodulator.

## FSK modulator/demodulator

The FSK transmitter frequency modulates the analogue output signal using two discrete frequencies to represent the binary data. The Bell 103 standard frequencies of 1270 Hz and 1070 Hz (originate mark and space) and 2225 Hz and 2025 Hz (answer mark and space), or the V21 standard frequencies of 980 Hz and 1180 Hz (originate mark and space) and 1650 Hz and 1850 Hz (answer mark and space) are used when this mode is selected. Demodulation involves detecting the received frequencies and decoding them into the appropriate binary value.

## Passband filters and equalisers

A bandsplit filter is included to shape the amplitude and phase response of the transmit signal into a square root, 75%-raised, cosine, and provided rejection of out-of-band signals in the receive channel.

## Asynchronous modes

The asynchronous mode is used for communication between asynchronous termi-

SPEED/PROTOCOL COMPATIBLILTY GUIDE

| | | | 73D2404 originating as: | | | | |
|---|---|---|---|---|---|---|---|
| | | | Bell | | CCITT | | |
| Calling a: | | | 300 | 1200 | 300 | 1200 | 2400 |
| Bell | 300 | (103) | 300 | 300 | - | - | 300 |
| | 1200 | (212) | 300 | 1200 | - | 1200 | 1200 |
| | 2400[1] | (224) | 300 | 1200 | - | 1200 | 2400 |
| CCITT | 300 | (V.21) | - | - | 300 | - | - |
| | 1200 | (V.22) | 300 | 1200 | - | 1200 | 1200 |
| | 2400 | (V.22 bis) | 300 | 1200 | - | 1200 | 2400 |
| | | | **73D2404 answering as:** | | | | |
| | | | **Bell** | | **CCITT** | | |
| **Called from a:** | | | **300** | **1200** | **300** | **1200** | **2400** |
| Bell | 300 | (103) | 300 | 300 | - | - | 300 |
| | 1200 | (212) | 300 | 1200 | - | 1200 | 1200 |
| | 2400 | (224) | 300 | 1200 | - | 1200 | 2400 |
| CCITT | 300 | (V.21) | - | - | 300 | - | - |
| | 1200 | (V.22) | 300 | 1200 | - | 1200 | 1200 |
| | 2400 | (V.22 bis) | 300 | 1200 | - | 1200 | 2400 |

[1] A Bell 2400 is a V.22 bis using a 2225 Hz answer tone without unscrambled marks.

890122 - T1

"AT" COMMANDS SUPPORTED
(Note: s=string; n=decimal, 0-255; x=boolean, 0/1=false/true)

| COMMAND | OPTIONS | DEFAULT |
|---|---|---|
| A/ | Repeats last command line | N/A |
| A | Answer | N/A |
| Bx | BELL/CCITT = 1/0 answer tone @1200 (N/A @2400) | 1 |
| Ds | Dial string specified by s | No string |
| Ex | Command echo, 0/1 = off/on | 1 |
| Hn | Hook status, 0/1 = on/off | N/A |
| In | ID code, 0/1/2/3/4 (see Table 8) | N/A |
| Kn | SSi test | N/A |
| Ln | Speaker volume, (0)1/2/3 = lo/med/hi | 2 |
| Mn | Speaker, 0/1/2/3 = control (see Table 3) | 1 |
| On | Online, 0/1/2/3 = on-line/retrain/no retrain (see Table 4) | N/A |
| P | Pulse dial | Pulse |
| Qx | Quiet result, 0/1 = 1-quiet | 0 |
| R | Reverse originate | N/A |
| Sn=n | Set S register (see Table 2) | N/A |
| Sn? | Return value in register n (see Table 2) | N/A |
| T | Touch tone dial | Pulse |
| Ux | User help screen, Sreg, dial string, data format | N/A |
| Vx | Verbose result, 0/1 = off/on | 1 |
| Xn | Result code, 0/1/2/3/4 (see Table 1) | 4 |
| Yx | Enable long space disconnect, 1 = enable | 0 |
| Z | Restore from Non-Volatile Memory | N/A |
| &Cx | Carrier detect override, 0/1 = on/normal | 0 |
| &Dn | DTR mode, 0/1/2/3 (see Table 5) | 0 |
| &F | Restore to factory configuration | N/A |
| &Gn | CCITT guard tone, 0/1/2 = off/1800/550 | 0 |
| &Jx | Auxiliary relay control | 0 |
| &Mn | Async/Sync mode, 0/1/2/3 (see Table 6) | 0 |
| &Rx | Enable RTS/CTS | 0 |
| &Sx | DSR override, 0/1=on/normal | 0 |
| &Tn | Test mode (see Table 7) | N/A |
| &Px | Pulse dial mode, 0/1=U.S./U.K. | 0 |
| &W | Write current configuration to NVRAM | N/A |
| &Xn | Sync Tx clock mode, 0/1/2=int/ext/slave | 0 |
| &Zs | Store a telephone number=string | N/A |

Factory configuration[1]:
B1 E1 F1 L2 M1 P Q0 V1 X4 Y0 &C0 &D0 &G0 &J0 &M0 &P0 &R0 &S0 &T4 &X0

Dial string arguments:
, = delay    @ = silent answer    ! = flash
; = return to command    s = dial stored number    W = wait for tone    R=reverse mode

[1] If the NovRAM has not been initialized it may be necessary to Power down/Power up and type AT&F&W<cr> to properly initialize modem state.

890122 - T2

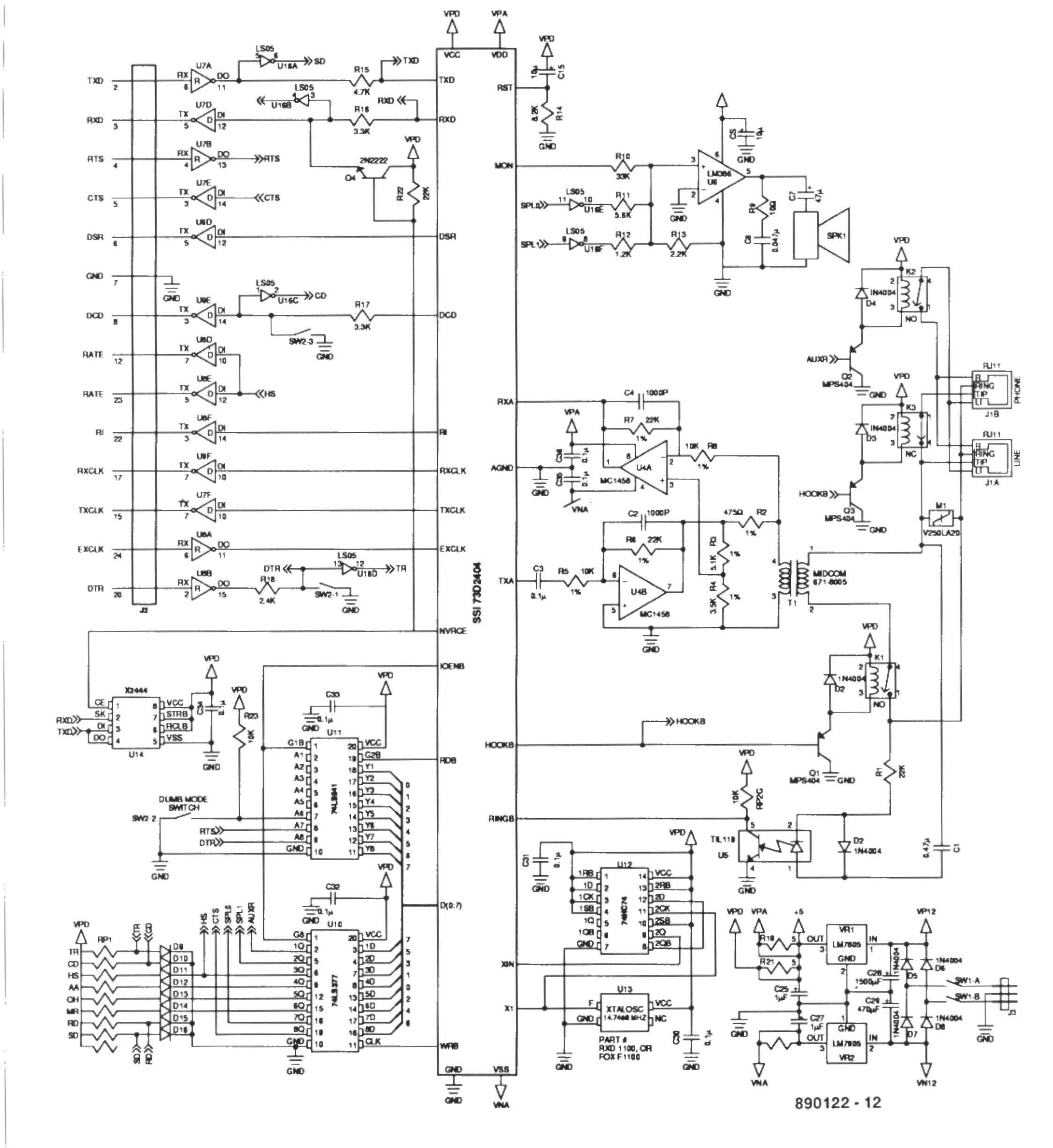



**Fig. 1. SSI 73D2404 modem chip set in a sample application circuit.**

nals which may vary the data rate from +1.5% to −1.5%. When transmitting in this mode, the serial data on the TXD input is passed through a rate converter that inserts or deletes stop bits in the serial bit stream in order to supply a signal whose data rate is accurate to 0.01%. The signal is routed to a data scrambler (following the CCITT V22bis algorithm) and into the modulator. The 73D2404 recognizes a break signal and handles it in accordance with Bell 212A specifications. Received data is processed in a similar manner except that the rate converter now acts to reinsert any deleted stop bits. An incoming break signal will be passed through without incorrectly inserting a stop bit.

## Synchronous modes

Synchronous operation is possible only with the QAM or PSK mode. Operation is similar to that of the asynchronous mode except that data must be synchronized to a clock, and no variation in data transfer rate is allowable. Serial input data appearing at the TXD terminal must

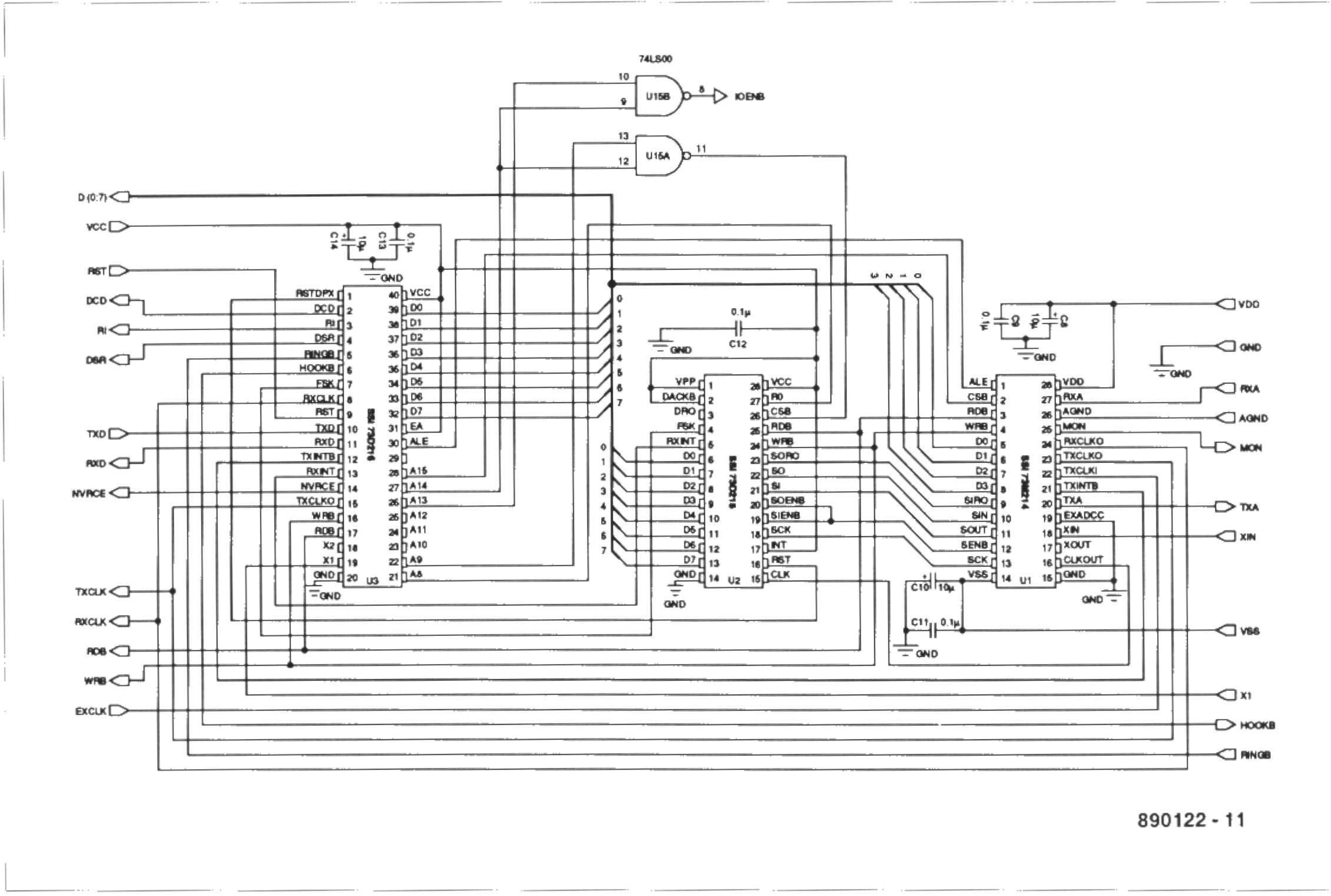



**Fig. 2. SSI73D2404 system interconnection diagram.**

be valid on the falling edge of signal TXCLK. Receive data at the RXD output is clocked out on the rising edge of signal RXCLK. The async/sync converter is bypassed when synchronous mode is selected and data is transmitted out at the same rate as is input.

## Automatic handshake

The SSI73D2404 will automatically perform a complete handshake as defined by V22bis, V22 and Bell 212A/103 standards to connect with a remote modem. The 73D2404 automatically determines the speed and operating mode and adjusts its operation to correspond to that of an answering modem when originating a call.

## Test modes

The SSI73D2404 allows the use of analogue loopback, digital loopback and remote digital loopback test modes. Full test mode capability allows testing of the modem and interface functions from the local terminal using the appropriate control commands, or remotely using the RDL function.

## Adaptive equalization with auto-retrain

The SSI73D2404 uses adaptive equalization which automatically compensates for varying line characteristics by adjusting taps on a multi-tap FIR filter. Optimum performance is obtained with this technique over a wide range of line conditions. When the line quality deteriorates to a specified level, the 73D2404 can automatically initiate a retrain of the equalizer to re-establish data communications without the need to go through a complete handshake sequence.

## AT-standard command interpreter

The SSI73D2404 includes an AT command interpreter that is compatible with the Hayes 2400 SmartModem™ command set. Functions and features included with intelligent modems are provided by the 73D2404 command interpreter. An overview of supported commands is given in Table 1.

## Non-volatile memory

The SSI73D2404 supports connection to an external non-volatile memory, e.g., Xicor's X2444, to store a dial string and the current AT command configuration. The use of a non-volatile RAM is illustrated in the application circuit.

**Source:**
*SSI73D2404 V.22bis 2400 bps Modem Device Set*, June 1988, published by Silicon Systems Inc.

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# PRACTICAL FILTER DESIGN – PART 8

by H. Baggott

**The Chebishev section has one of the steepest cut-off profiles of all types of filter. Unfortunately, it also has a limiting deficiency: a ripple in the pass band. The Chebishev filter can be dimensioned in various ways: the ripple is at all times limited to a certain value. This part of the series includes the Chebishev design tables for a ripple of 0.1 dB.**

The Chebishev function is one of the most effective functions for realizing a filter: it combines a pronounced bend at the cut-off point with a sharp profile. This combination also results in ringing, which, by careful design can fortunately be kept within a given value. There is, however, a direct relation between the cut-off profile and the ringing: if the former is made steeper, the latter becomes more pronounced; and if the ringing is kept to a small value, the profile is less steep. In practice, a compromise is reached, be-cause in virtually all applications a ripple exceeding 1 dB is unacceptable. This part and Part 9 will deal with Chebishev filters with a 0.1 dB and a 0.5 dB ripple respectively. These are the values that satisfy most applications.

A general drawback of Chebishev filters is the very irregular delay time that, for instance, makes the filter unsuitable for use in loudspeaker cross-over networks.

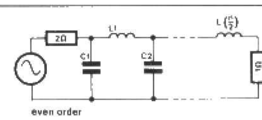
The computation of the Chebishev poles can be done in two ways. In the first, use is made of the Chebishev polynomials, while in the second the real part of the poles of a Butterworth transfer function are multiplied with a constant factor, which results in a shifting of the poles from a circle to an ellipse. Note that in the Chebishev polynomials the cut-off point is not at –3 dB, but the tables take this into account.

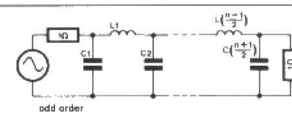
Table 10

| n | real part $-\alpha$ | imaginary part $\pm\beta$ |
|---|---|---|
| 2 | 0.6074 | 0.7112 |
| 3 | 0.348 | 0.871 |
| | 0.696 | |
| 4 | 0.2174 | 0.9292 |
| | 0.5248 | 0.3849 |
| 5 | 0.1466 | 0.9565 |
| | 0.3838 | 0.5912 |
| | 0.4744 | |
| 6 | 0.1049 | 0.9715 |
| | 0.2865 | 0.7112 |
| | 0.3913 | 0.2603 |
| 7 | 0.07846 | 0.9806 |
| | 0.2198 | 0.7863 |
| | 0.3177 | 0.4364 |
| | 0.3526 | |
| 8 | 0.06079 | 0.9864 |
| | 0.1731 | 0.8363 |
| | 0.2591 | 0.5588 |
| | 0.3056 | 0.1962 |
| 9 | 0.04844 | 0.9905 |
| | 0.1395 | 0.871 |
| | 0.2137 | 0.6465 |
| | 0.2621 | 0.344 |
| | 0.2789 | |
| 10 | 0.03947 | 0.9934 |
| | 0.1145 | 0.8962 |
| | 0.1784 | 0.7112 |
| | 0.2248 | 0.4566 |
| | 0.2492 | 0.1573 |

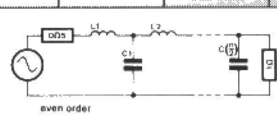
**Table 10. Pole locations of Chebishev filters with 0.1 dB ripple.**

Table 11




| n | C1 | L1 | C2 | L2 | C3 | L3 | C4 | L4 | C5 | L5 |
|---|---|---|---|---|---|---|---|---|---|---|
| 2 | 0.08908 | 0.486 | | | | | | | | |
| 3 | 0.228 | 0.2536 | 0.288 | | | | | | | |
| 4 | 0.06999 | 0.5136 | 0.1539 | 0.4546 | | | | | | |
| 5 | 0.2071 | 0.2476 | 0.3567 | 0.2476 | 0.2071 | | | | | |
| 6 | 0.06584 | 0.4883 | 0.1524 | 0.5908 | 0.1559 | 0.4446 | | | | |
| 7 | 0.2008 | 0.2419 | 0.3564 | 0.2674 | 0.3564 | 0.2419 | 0.2008 | | | |
| 8 | 0.06433 | 0.4779 | 0.1498 | 0.5876 | 0.161 | 0.5987 | 0.1554 | 0.4407 | | |
| 9 | 0.1981 | 0.239 | 0.3536 | 0.2678 | 0.3654 | 0.2678 | 0.3536 | 0.239 | 0.1981 | |
| 10 | 0.06363 | 0.4729 | 0.1484 | 0.5833 | 0.1606 | 0.6036 | 0.1626 | 0.5995 | 0.155 | 0.4388 |
| | L1 | C1 | L2 | C2 | L3 | C3 | L4 | C4 | L5 | C5 |



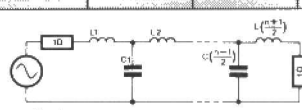
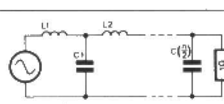
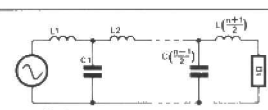




**Table 11. Standardized component values for passive low-pass filters with an input impedance to output impedance ratio of 2:1 for even order sections and 1:1 for odd-order filters.**

Table 12




| n | L1 | C1 | L2 | C2 | L3 | C3 | L4 | C4 | L5 | C5 |
|---|---|---|---|---|---|---|---|---|---|---|
| 2 | 0.2214 | 0.1304 | | | | | | | | |
| 3 | 0.2408 | 0.2402 | 0.114 | | | | | | | |
| 4 | 0.2404 | 0.2814 | 0.2316 | 0.107 | | | | | | |
| 5 | 0.2485 | 0.2876 | 0.2811 | 0.2256 | 0.1036 | | | | | |
| 6 | 0.2441 | 0.2998 | 0.2913 | 0.2783 | 0.2218 | 0.1016 | | | | |
| 7 | 0.2506 | 0.2957 | 0.3057 | 0.2908 | 0.276 | 0.2914 | 0.1004 | | | |
| 8 | 0.2454 | 0.3041 | 0.3025 | 0.3064 | 0.2897 | 0.2742 | 0.2178 | 0.0996 | | |
| 9 | 0.2515 | 0.298 | 0.3117 | 0.3039 | 0.306 | 0.2886 | 0.273 | 0.2166 | 0.09904 | |
| 10 | 0.2461 | 0.3056 | 0.3058 | 0.3135 | 0.304 | 0.3055 | 0.2878 | 0.272 | 0.2158 | 0.09864 |

**Table 12. Standardized component values for passive low-pass sections with negligible source impedance.**

Table 13

| n | C1 | C2 | C1 | C2 | C3 |
|---|---|---|---|---|---|
| 2 | 0.2607 | 0.1107 | | | |
| 3 | | | 1,0589 | 0.2905 | 0.02141 |
| 4 | 0.7308 | 0.03836 | | | |
| | 0.3024 | 0.1975 | | | |
| 5 | 1,0838 | 0.02515 | | | |
| | | | 0.7076 | 0.4011 | 0.06054 |
| 6 | 1,5169 | 0.01767 | | | |
| | 0.555 | 0.07826 | | | |
| | 0.4063 | 0.2827 | | | |
| 7 | 2,026 | 0.01304 | | | |
| | 0.7235 | 0.05301 | | | |
| | | | 0.8236 | 0.5287 | 0.09061 |
| 8 | 2,6165 | 0.01001 | | | |
| | 0.9188 | 0.03817 | | | |
| | 0.6139 | 0.1097 | | | |
| | 0.3613 | 0.3697 | | | |
| 9 | 3,285 | 0.00793 | | | |
| | 1,1411 | 0.02884 | | | |
| | 0.7445 | 0.07409 | | | |
| | | | 0.9858 | 0.6622 | 0.1191 |
| 10 | 4,0298 | 0.00643 | | | |
| | 1,3889 | 0.02258 | | | |
| | 0.8917 | 0.0534 | | | |
| | 0.7078 | 0.1394 | | | |
| | 0.6384 | 0.4579 | | | |

**Table 13. Standardized component values for active filters with single feedback path.**

## Filter with 0.1 dB ripple

Tables 10–14 give all information necessary for the computation of Chebishev filters from the 2nd to the 10th order with a ripple of 0.1 dB. Note, however, that the values for odd-order filters in Table 11 apply to sections whose input to output impedance ratio is 1:2 (if a T configuration) or 2:1 (if a $\pi$ filter).

The gain vs frequency curves in Fig. 42 clearly show the sharp profile of this type of filter, while the ripple is hardly noticeable. The delay time vs frequency curves in Fig. 43 give a rather worse picture than those of the filters discussed in previous parts of this article. The step response in Fig. 44 clearly shows the ringing. Note that the curves in Fig. 43 and Fig. 44 do not improve all that much if a lower value ripple is chosen.

## Two examples

As in previous parts, we give two examples of how to compute a filter. This time, we take a band-pass filter and a complex low-pass section in a double opamp configuration.

### Example 1.

Compute a passive band-pass filter with a centre frequency of 1 kHz and a bandwidth of 100 Hz. The attenuation at 900 Hz and 1100 Hz must be at least 20 dB. The output impedance is to be 600 $\Omega$ and the output impedance of the amplifier to which the filter is to be connected is negligible.

Solution.

Since the centre frequency is known, it need not be computed. We calculate the frequencies corresponding to 900 Hz and 1100 Hz but at the opposite side of the filter referred to the centre frequency to find the sharpest roll-off. The frequency associated with 900 Hz is:

$f_2 = 1000^2 / 900 = 1111$ Hz,

and that associated with 1100 Hz is:

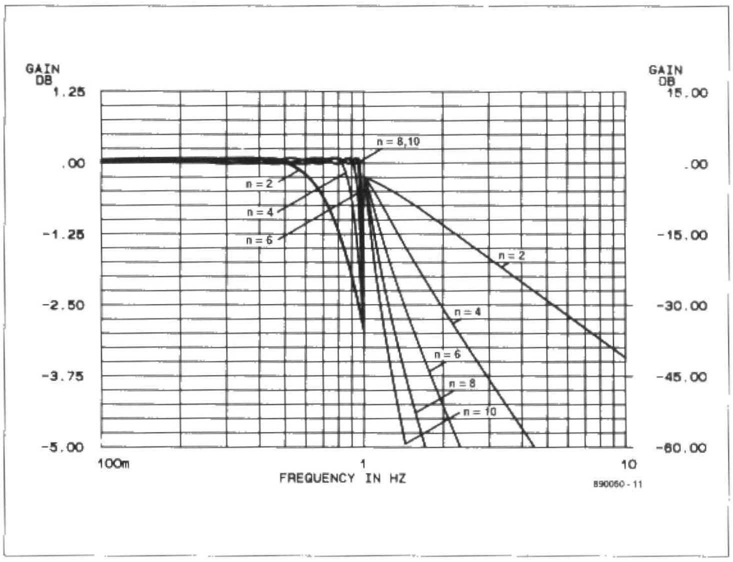



**Fig. 42. Gain vs frequency characteristics of Chebishev filters with a 0.1 dB ripple.**

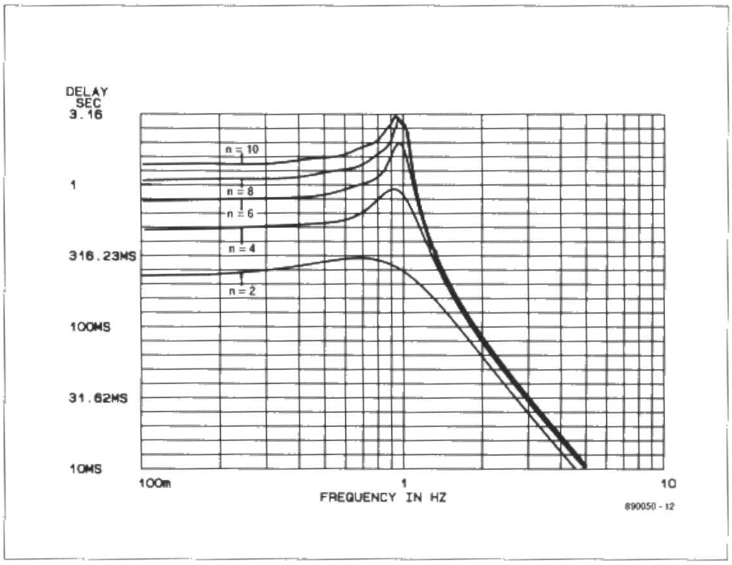



**Fig. 43. Delay time vs frequency characteristics of Chebishev filters with a 0.1 dB ripple.**

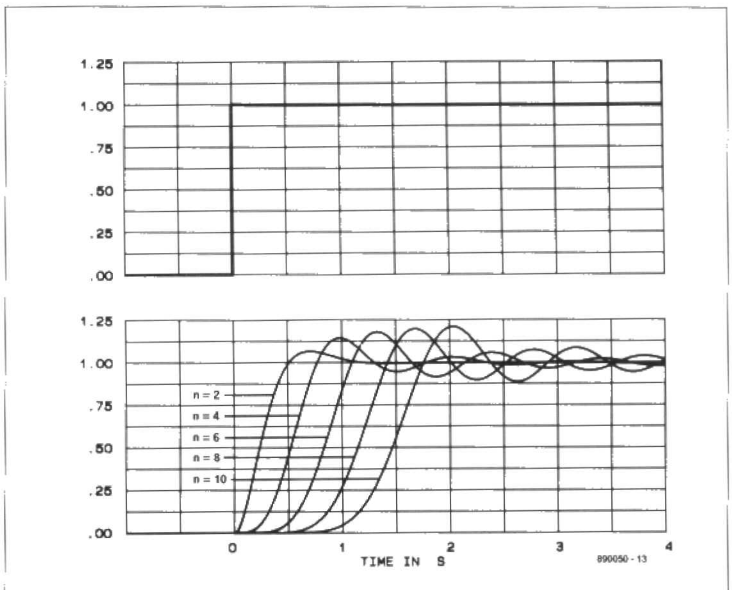



**Fig. 44. Step response of Chebishev filters with a 0.1 dB ripple.**

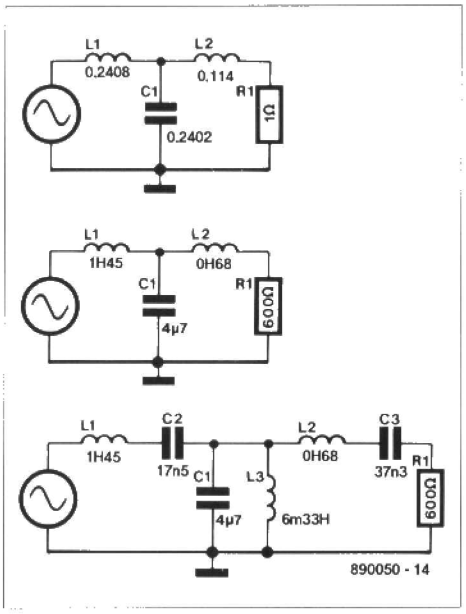



**Fig. 45. Designing a passive band-pass filter: (a) standardized high-pass section; adjusting values for required bandwidth; (c) conversion to band-pass filter**

$f_1 = 1000^2 / 1100 = 909$ Hz.

The frequencies closest to 1 kHz are 909 Hz and 1100 Hz, so that the bandwidth at the –20 dB points is 191 Hz.

In the characteristics, we now have to find a filter that provides an attenuation of not less than 20 dB at a standardized frequency of 191/100 = 1.91 Hz (note that for a low-pass section the bandwidth, not the central frequency, is the basis of the design). From the data, we choose a third-order, 0.1 dB Chebishev section: this provides an attenuation of about 22 dB at $f = 2$ Hz (estimated between a second- and a fourth-order filter).

Figure 45a shows the circuit diagram of a third-order low-pass section. The standardized component values are derived from Table 12. Next, the 'real' values are calculated for a terminal impedance of 600 Ω and a cut-off frequency equal to the –3 dB bandwidth (100 Hz).

$L_1 = LR/f = 1.4448$ H

$C_1 = C/Rf = 4.003\times10^{-6} = 4\ \mu$F

$L_2 = LR/f = 0.684$ H

Then follows the transformation from a low-pass to a band-pass filter as explained in Part 5, which results in the filter shown in Fig. 45c. The remaining components are calculated with the aid of formulas [34] and [35] (Part 5):

$C_2 = 1/L(2\pi f_c)^2 = 1/1.45(2\pi1000)^2 =$

$= 1.75\times10^{-8} = 17.5$ nF

$L_3 = 1/C_1(2\pi f_c)^2 = 1/4\times10^{-6}(2\pi1000)^2 =$

$= 6.33\times10^{-3} = 6.33$ mH

$C_3 = 1/L_2(2\pi f_c)^2 = 1/0.68(2\pi1000)^2 =$

$= 3.73\times10^{-8} = 37.3$ nF

Note that the central frequency of the band-pass section is used only for the computation of the values of those components that are added during the conversion process.

**Example 2.**

Design an active high-pass filter with a cut-off frequency of 3 kHz and a slope of 12 dB/octave. It is essential that the cut-off frequency can be set accurately. The gain of the section must be 14 dB (×5).

Solution.

This type of section is best realized by a state-variable filter (see Fig. 17 – Part 3). For convenience, we again choose a Chebishev filter. The state-variable filter is based on the poles of a second-order filter in Table 10:

$-\alpha = 0.6074$

$\beta = \pm0.7112$

First, we choose a value for $C$, say, 4.7 nF. Resistors $R$ are given a value of 33 kΩ. The other resistor values are then calculated with the aid of formulas [19], [20], [21] and [22], but note that all values so obtained must be divided by the cut-off frequency since the formulas give values for $f = 1$ Hz.

$R_1 = 1/[2\pi f_k AC\sqrt{(\alpha^2+\beta^2)}] =$
$= 1/2\pi\times3000\times5\times4.7\times10^{-9}\times$
$\times\sqrt{(0.6074^2+0.7112^2)} = 2414\ \Omega$

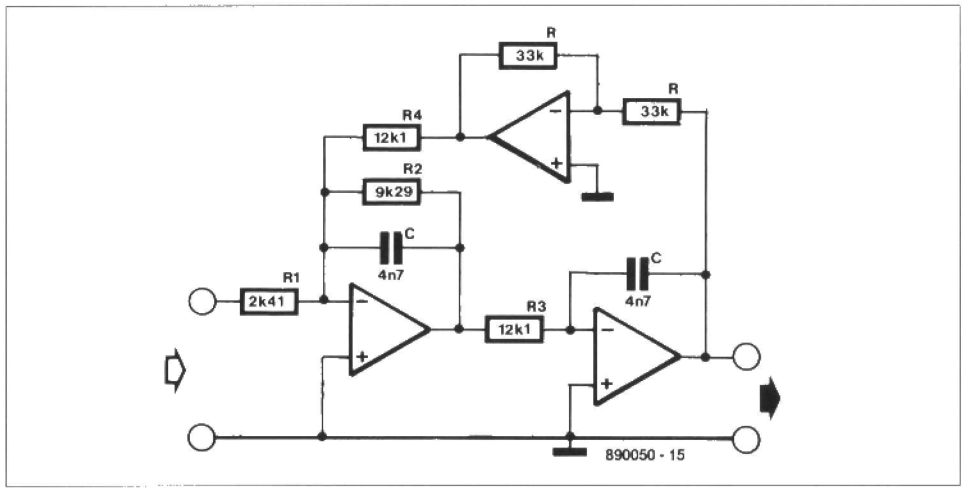



**Fig. 46. Example of a state-variable filter with a cut-off frequency of 3 kHz.**

$R_2 = 1/4\pi\alpha Cf_k =$
$= 1/4\pi\times3000\times0.6074\times4.7\times10^{-9} =$
$= 9291\ \Omega$

$R_3 = R_4 = 1/2\pi Cf_k\sqrt{(\alpha^2+\beta^2)} =$
$= 1/2\pi\times3000\times4.7\times10^{-9}\times$
$\times(0.6074^2+0.7112^2) = 12068\ \Omega$

Resistors $R_2$ and $R_4$ may be a combination of a fixed resistor and a preset potentiometer to enable the cut-off frequency and Q-factor of the section to be set accurately.

## Correction to Part 3

From the foregoing in this part of the series, you will have noticed that resistors $R_2$ and $R_4$ and NOT $R_1$ and $R_3$ (as stated in Part 3) are used for setting the parameters of the state-variable section. Thus, $R_4$ serves to set the maximum output voltage of $A_1$ at $f_o$, while $R_2$ is used to set the bandwidth to the value at which the Q-factor is calculated.

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| Jumbo clock 2 × 2716 | 539 | 14.60 | 2.20 |
| Graphics card 2 × 82S123 | 543 | 9.60 | 1.44 |
| printer buffer 1 × 2716 | 545 | 7.30 | 1.10 |
| MSX EPROMmer 1 × 27128 | 552UK | 7.30 | 1.10 |
| Intelligent time standard 1 × 2764 | 553 | 10.00 | 1.50 |
| Interface for high-resolution LC screens 1 × 2764 | 560 | 10.00 | 1.50 |
| I/O extension for IBM 1 × PAL16L8 | 561 | 8.75 | 1.32 |
| Centronics interface for slide fader 1 × PAL16R4 | 562 | 8.75 | 1.32 |
| µP-controlled radio synthesizer 1 × 27C64 | 564 | 10.00 | 1.50 |
| Portable MIDI keyboard 1 × 2764 | 567 | 10.00 | 1.50 |
| Pitch control for CD players 1 × 2764 | 568 | 10.00 | 1.50 |
| MIDI control unit 1 × 27C64 | 570 | 10.00 | 1.50 |
| Slave indication unit for I.T.S. 1 × 8748H | 700 | 15.00 | 2.25 |
| EPROM emulator 1 × 8748H | 701 | 15.00 | 2.25 |
| Microcontroller-driven power supply 1 × 8751 | 702 | 47.50 | 7.13 |
| Autonomous I/O controller 1 × 8751 | 704 | 47.50 | 7.13 |
| Transmission & reception of RTTY | XSS-100 | 8.00 | 1.20 |
| (Send formatted 3.5 in diskette containing MSXDOS.COM and COMMAND.COM) | | | |
| FAX interface for Atari ST | ESS-102 | 8.00 | 1.20 |
| for Archimedes | ESS-103 | 8.00 | 1.20 |
| (Send formatted 3.5 in. diskette) | | | |

### PRINTED CIRCUITS

**Readers who wish to make their own PCBs (for private and personal use only) can order the relevant drawings from our editorial offices in London, quoting the relevant PCB number and indicating whether they are required on paper or on film. Prices per drawing, including postage & packing, are £1.50 (paper — surface mail), £2.00 (paper - airmail outside Europe), £5.50 (film — surface mail), or £7.00 (airmail outise Europe).**

| | No. | Price (£) | VAT (£) |
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| **DECEMBER 1988** | | | |
| LFA-150: a fast power amplifier | 880092-3 | 6.40 | 0.96 |
| | 890092-4 | 6.45 | 0.97 |
| Composite-to-TTL adaptor | 880098 | 4.85 | 0.73 |
| Colour test pattern generator | 880130 | 13.30 | 2.00 |
| Autonomous I/O controller | 880162 | 4.50 | 0.68 |
| | 880163 | 4.85 | 0.73 |
| | 880184 | 15.30 | 2.30 |
| Pitch control for CD players | 880165 | 11.50 | 1.73 |
| **JANUARY 1989** | | | |
| Thyristor speed control | 87200 | Not available | |
| Fax interface for Atari ST/Archimedes | 880109 | 7.35 | 1.10 |
| MIDI control unit | 880178-1 | 9.05 | 1.36 |
| | 880178-2 | 6.65 | 1.00 |
| Low-budget capacitance meter | UPB-S1 | 1.95 | 0.29 |
| **FEBRUARY 1989** | | | |
| MOSFET power amplifier | 87096 | 12.35 | 1.85 |
| The digital train | 87291-1 | 4.20 | 0.63 |
| Touch key organ | 886077 | 10.05 | 1.16 |
| Car service module | 86765 | 3.60 | 0.54 |
| | 886126 | 4.10 | 0.62 |
| VHF receiver | 886127 | 7.45 | 1.12 |
| Dark-room timer | 886100 | Not available | |
| **MARCH 1989** | | | |
| Diesel sound generator | 880006 | Not available | |
| VHF/UHF wide-band amplifiers | 880186 | Not available | |
| Power line modem | 880189 | 6.10 | 0.92 |
| ATN Filmnet decoder | 890002 | Not available | |
| Centronics buffer | 890007-1 | 19.60 | 2.94 |
| | 890007-2 | 2.15 | 0.33 |
| | 890007-3 | 8.35 | 1.26 |

| | No. | Price (£) | VAT (£) |
|---|---|---|---|
| **APRIL 1989** | | | |
| The digital train | 87291-2/3 | 4.30 | 0.65 |
| Class D amplifier | 880009 | Not available | |
| DLF frequency reference | 880197 | Not available | |
| Function generator | UPBS-1 | 1.95 | 0.29 |
| Triplet | 890013-1 | 6.65 | 1.00 |
| | 890013-2 | 6.80 | 1.02 |
| Multi-point IR control | 890019-1 | 3.45 | 0.52 |
| | 890019-2 | 4.05 | 0.61 |
| Video recording amplifier | 889502 | See advert on pp.8-9 | |
| **MAY 1989** | | | |
| RDS decoder | 880209 | Not available | |
| Digital model train (4) | 87291-4 | 5.25 | 0.79 |
| Analogue Multimeter | 890035 | 12.50 | 1.88 |
| In-line RS-232 monitor | 890036 | Not available | |
| Code converter for Centronics-compatible printers | 890058 | Not available | |
| DTMF system decoder | 890060 | 6.50 | 0.98 |
| Transistor tester | 896029 | Not available | |
| Sine wave converter | UPB-S1 | 1.95 | 0.29 |
| Super-VHS to RGB converter | See advert on pp. 8-9 | | |
| **JUNE 1989** | | | |
| 8-digit frequency meter | 880128 | 11.50 | 1.73 |
| In-line RS232 monitor | 890036 | Not available | |
| In-circuit transistor tester | 890079 | Not available | |
| Echo unit | See advert on p.8-9 | | |
| **July/August 1989** | | | |
| MIDI keyboard interface | | | |
| decoder board | 890105-1 | 7.00 | 1.05 |
| controller board | 890105-2 | 5.25 | 0.79 |
| Tracking tester | See advert on pp. 8-9 | | |
| Test pattern generator | 890020 | Not available | |
| DC-AC power converter | 890056 | Not available | |
| Floppy disk monitor | 890078 | 4.25 | 0.64 |
| display board | 85019 | 3.25 | 0.48 |
| Function generator | UPBS-1 | 1.95 | 0.29 |
| Call tone generator | 894016 | Not available | |
| Sound level meter | 894024 | 3.50 | 0.53 |
| Mains-failure indicator | 894030 | Not available | |
| Radio beacon converter | 894041 | Not available | |
| Low-noise microphone preamplifier | 894063 | 3.75 | 0.56 |
| Bucket brigade delay line | 894055 | 3.75 | 0.56 |
| **SEPTEMBER 1989** | | | |
| Digital model train | 87291.IV | 6.70 | 1.00 |
| Stereo viewer | 890044 | 5.57 | 0.86 |
| Simple FM receiver | 890118 | Not available | |
| Centronics monitor | 890123 | 5.50 | 0.83 |
| Resonance meter | 886071 | 3.90 | 0.59 |

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